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POWER SYSTEM ANALYSIS

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Texas Tech University
Lubbock, TX 79409

January 1979

Final Report

Approved for public release; distribution unlimited.

AIR FORCE WEAPONS LABORATORY
Air Force Systems Command
Kirtland Air Force Base, NM 87117

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This report describes the system analysis performed on a lightweight, high-power power system. The system consists primarily of the turboalternator, rectifier/filter, and fault protection circuits. The determination of and effects from various alternator faults are described. Calculations for the filter weight and volume and a grounding and shielding scheme are presented. A design concept for a load protection crowbar is developed. Because a small, high voltage power supply is required for an electron-beam gun,

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BLOCK 20. ABSTRACT (CONT'D)

methods of generating high voltages in small packages are reviewed. Also included is a review of the state of the art of lightweight power sources, including advanced conventional, permanent magnet, and superconducting alternators, magnetohydrodynamic generators (MHD), and batteries.

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INTRODUCTION

The power system under consideration consists of a turboalternator with a coupling gearbox, rectifier/filter, and crowbar. In addition, a high voltage auxiliary supply is required by the load. Because the power system is to be lightweight, the designs of the individual components must incorporate high technology materials, and detailed analysis must be performed to insure airworthiness and reliability. Accordingly, analysis on the alternator as related to faults is an important consideration in determining stresses on the drive train. Engineering estimates of the magnitude of expected fault torques can be simply calculated using the law of conservation of flux linkages and the machine constants. Analyses of the turbine, gearbox, fluid system, etc. are beyond the scope of this report.

Because the load requires D.C. power, a rectifier/filter combination is necessary to condition the A.C. output of the alternator. Several designs are possible for the filter components, all of which result in lightweight units. Because of possible interference of the magnetic field generated by a filter inductor, an air core, toroidal design is chosen for analysis. Also, because of the limited duty cycle for the power system, a heat sink design is employed to further reduce the weight and complexity of the filter design. The rectifiers are silicon controlled rectifiers (SCRs) to facili-

tate on-off control of the load voltage and fault control. Because an optimum weight filter design usually exhibits a considerable overshoot for a snap-on condition, the SCRs can be controlled to phase on the power supply so that no over-voltage occurs at the load. However, analysis of the details of the SCR control is not covered in this report.

In order to prevent excessive damage to the load in the event of a load arc, a crowbar can be employed. By shorting the power supply immediately after a load fault occurs, energy can be diverted from the fault. A scheme for which the crowbar is slave triggered is the most reliable. Thus, the occurrence of a fault will automatically trigger the crowbar without an intervening control step. Because the crowbar circuit contains inductance, some energy cannot be diverted from the load by the crowbar. By using additional circuits, forced commutation of the load arc will rapidly extinguish the arc so that arc damage is held to a minimum.

The load requires an auxiliary power supply (E-beam supply), which imposes additional conditioning requirements. This supply requires relatively low power and high voltage. Because of weight and volume limitations, several schemes to achieve the required performance can be evaluated to determine the most likely candidates. The E-beam supply power input is restricted to the auxiliary power unit (APU). Therefore the unit must be electrically compatible with the main power supply. A description and comparison of simple transformer-rectifier, inverter-transformer,

voltage multiplier, and pulse charge circuits will serve as a reasonable cross section for E-beam power supply evaluation.

Grounding and shielding are important considerations for reliable system operation, especially where important control functions are performed at low levels in a high voltage/high power environment. A great many interference problems can be eliminated by careful grounding. A common point grounding scheme is usually employed to prevent interference arising because of ground loops, especially in the high power circuits. This scheme prevents cross talk between circuits because of a common ground return conductor. Shielding, when properly employed, can effectively eliminate interference due to electromagnetic (EM), fields. Because the expected spectrum of noise is rather low frequency, the shielding can be broken into a quasisteady state field analysis, i.e., electrostatic and magnetic. Magnetic shielding is handled by component design to eliminate or minimize stray fields and is not addressed as a separate topic in this report. Electrostatic shielding can be used to prevent circuit damage under fault conditions as well as prevent interference. By proper placement and grounding of the shields, fault currents as well as currents due to electrostatic pickup can be conducted to ground potential without interference to low level signal conductors.

Because of further power supply requirements, a review of the development of power sources is useful. A review of the

major problems and work performed toward their solution serves to compare the sources. The classical turboalternator, superconducting alternator, magnetohydrodynamic (MHD) generator, and storage batteries are the principal contenders.

MACHINE FAULTS

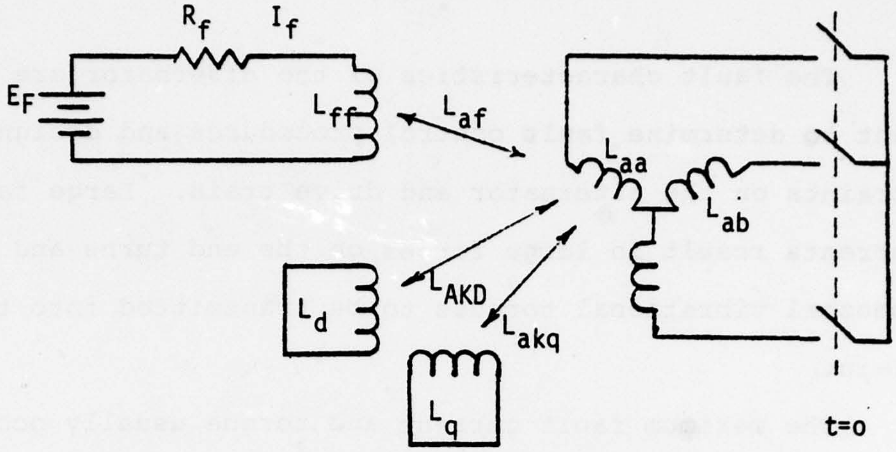
The fault characteristics of the alternator are important to determine fault control procedures and design constraints on the alternator and drive train. Large fault currents result in large forces on the end turns and cause abnormal vibrational torques to be transmitted into the drive train.

The maximum fault current and torque usually occur within the first cycle after the occurrence of the fault. Thus, the simplest method of calculating the initial character of the fault response involves using the law of conservation of flux linkages. By utilizing this law and the inductance matrix of the machine, the fault currents are easily obtained by computer solution. The fault torque is obtained by the relation

$$T = \sum i_p^2 \frac{\partial L_p}{\partial \theta} + \sum_{m \neq n} i_m i_n \frac{\partial L_{mn}}{\partial \theta} \qquad (1)$$

The inductance matrix used in these calculations is shown in Figure 1. Although saturation effects are not included, this procedure will serve to estimate the effects of fault conditions on the APU.

An estimate of fault torque can be obtained for the condition of a no-load to three phase, line-to-line fault with field excitation such that rated terminal voltage is achieved prior to the fault. Maximum fault currents occur for this condition because the net flux in the machine is greatest



$$
\begin{bmatrix} \psi_a \\ \psi_b \\ \psi_c \\ \psi_f \\ \psi_d \\ \psi_q \end{bmatrix}
=
\begin{bmatrix}
L_{aa} & L_{ab} & L_{ac} & L_{af} & L_{AKD} & L_{akq} \\
L_{ab} & L_{bb} & L_{bc} & L_{bf} & L_{bkd} & L_{bkq} \\
L_{ac} & L_{bc} & L_{cc} & L_{cf} & L_{ckd} & L_{ckq} \\
L_{af} & L_{bf} & L_{cf} & L_{ff} & L_{fkd} & 0 \\
L_{akd} & L_{bkd} & L_{ckd} & L_{fkd} & L_{dd} & 0 \\
L_{axq} & L_{bkq} & L_{cxq} & 0 & 0 & L_{qq}
\end{bmatrix}
\begin{bmatrix} i_a \\ i_b \\ i_c \\ i_f \\ i_d \\ iq \end{bmatrix}
$$

Figure 1. Machine model and inductance matrix

for no-load, rated voltage conditions. Under load conditions, there is an armature reaction flux which cancels a portion of the excitation flux, and as a result, there is a net reduction in total air gap flux even though the field excitation may be considerably larger than that required to produce rated terminal voltage under no-load conditions. Before the fault occurs, the only existing current is the field current. By noting the flux linking each winding just prior to the fault and by obtaining the values of flux linkages after the fault occurs, the solution of the required winding currents to maintain these flux linkages is straightforward.

The fault currents and torque for a field current of 200 amperes are shown in Figures 2 thru 7. The maximum phase current is shown to be approximately 3,800 amperes, and the direct axis damper winding has a peak current of 4,470 amperes. Phase A current has a large D.C. component because the fault occurs at the time when the field axis is aligned with the phase A axis. In Figure 3, it is seen that the peak air gap torque is 12,700 newton meters or 9,350 pound-feet. Because the rotor is not an ideal rigid body, not all of the torque is transmitted to the alternator shaft. A complete knowledge of the rotor is required to determine the exact character of the shaft torque. Assuming the worst case (all the air gap torque transmitted to the shaft), the gear box will be required to sustain approximately four times rated torque, vibrating at alternator frequency.

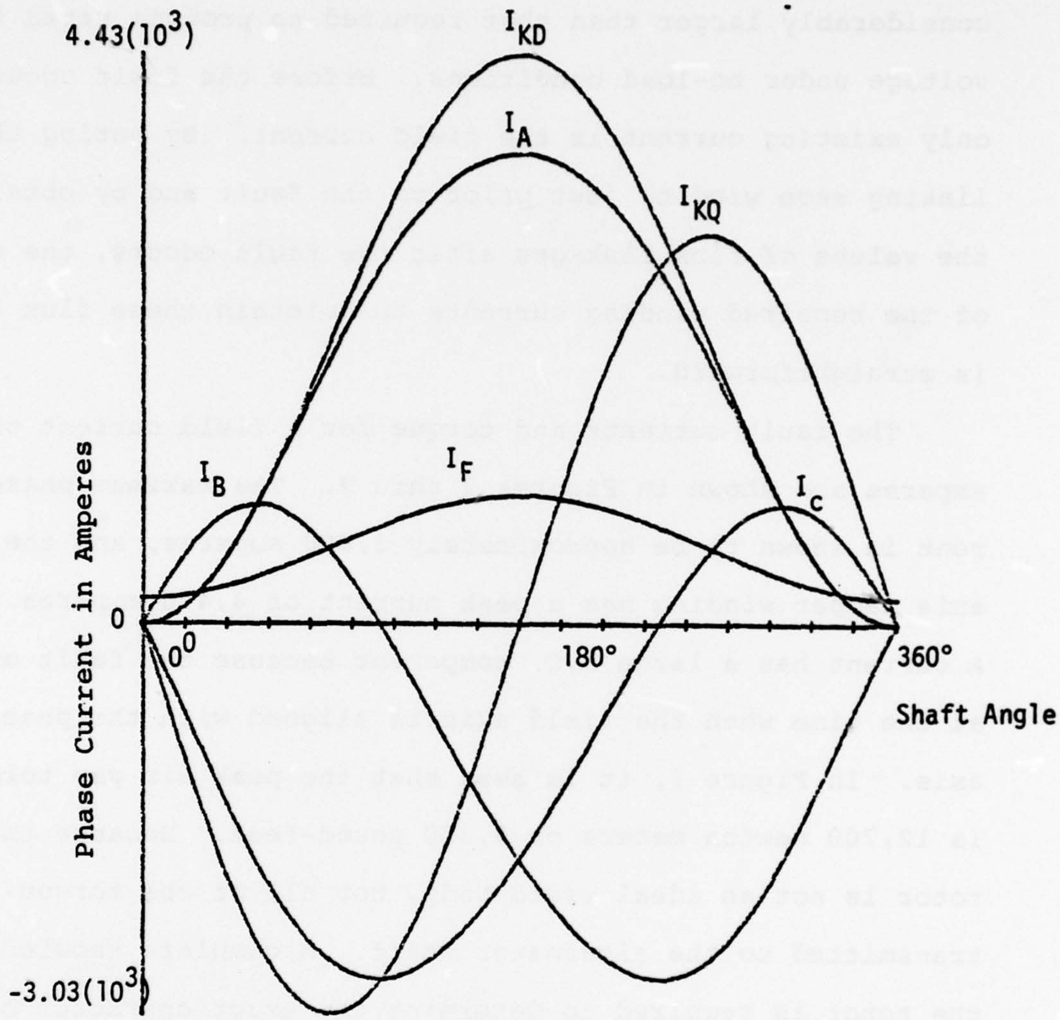


Figure 2. Three phase short; current response

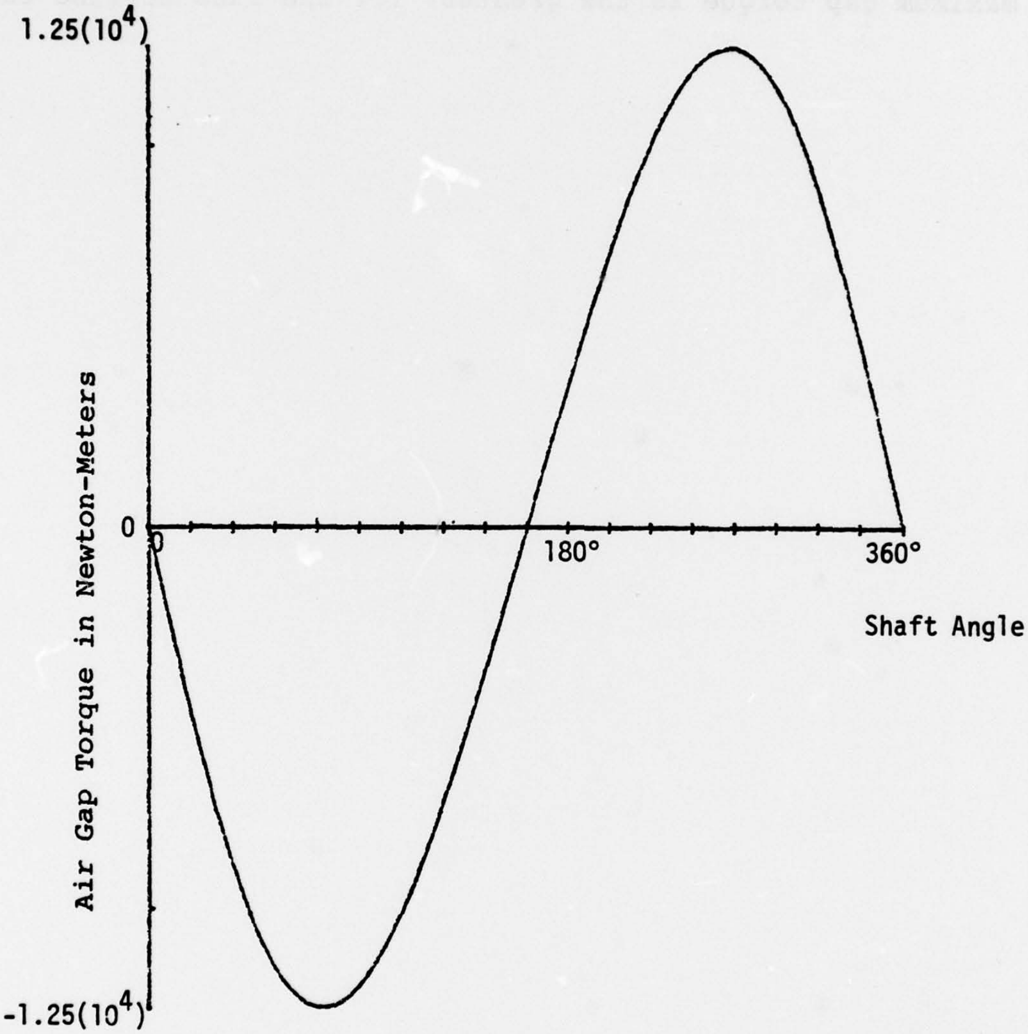


Figure 3. Three phase short; torque response

Figures 4 and 5 show the current and air gap torque response for a single phase to ground fault. Figures 6 and 7 show similar responses for a line to line fault. Note that of the three examples presented in Figure 2 thru 7, the maximum gap torque is the greatest for the line-to-line fault.

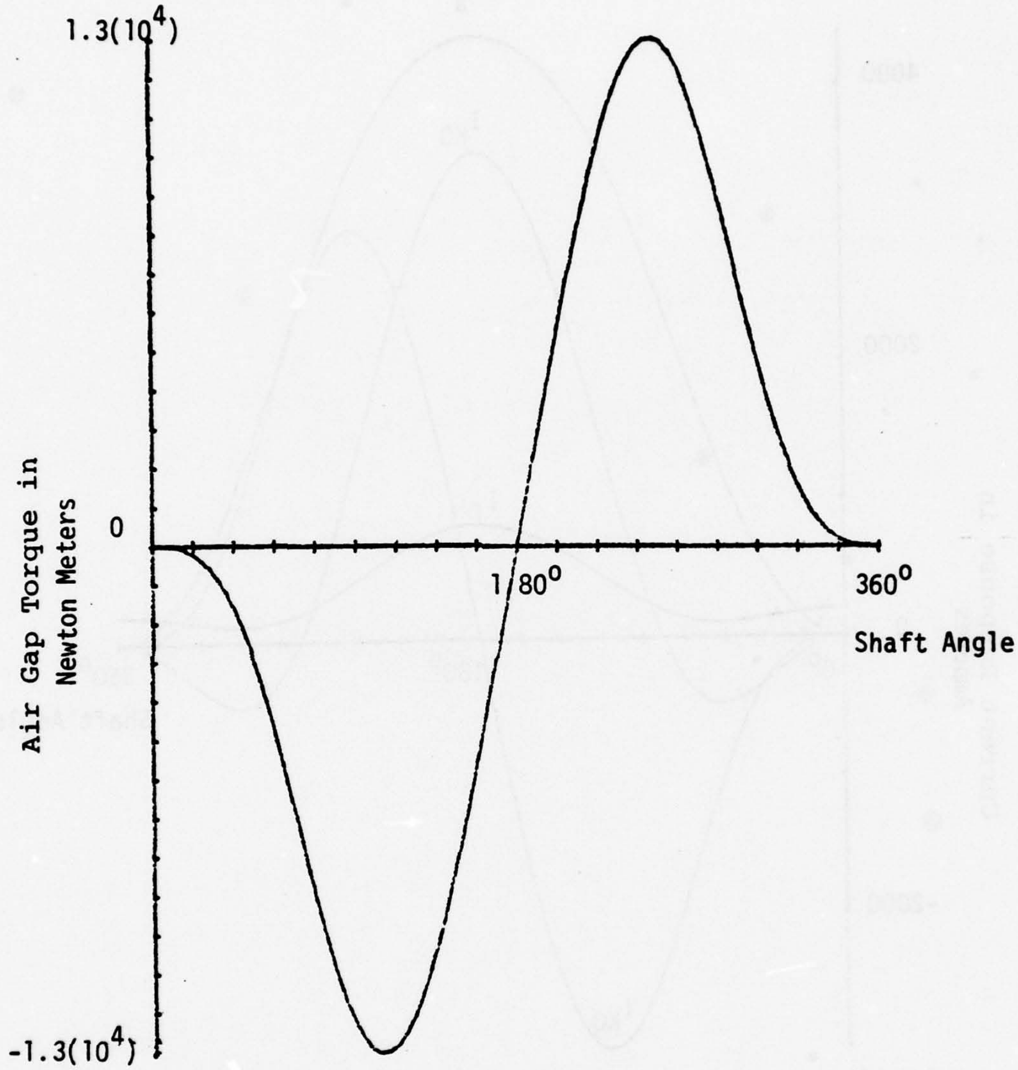


Figure 4. Phase A to ground fault torque response

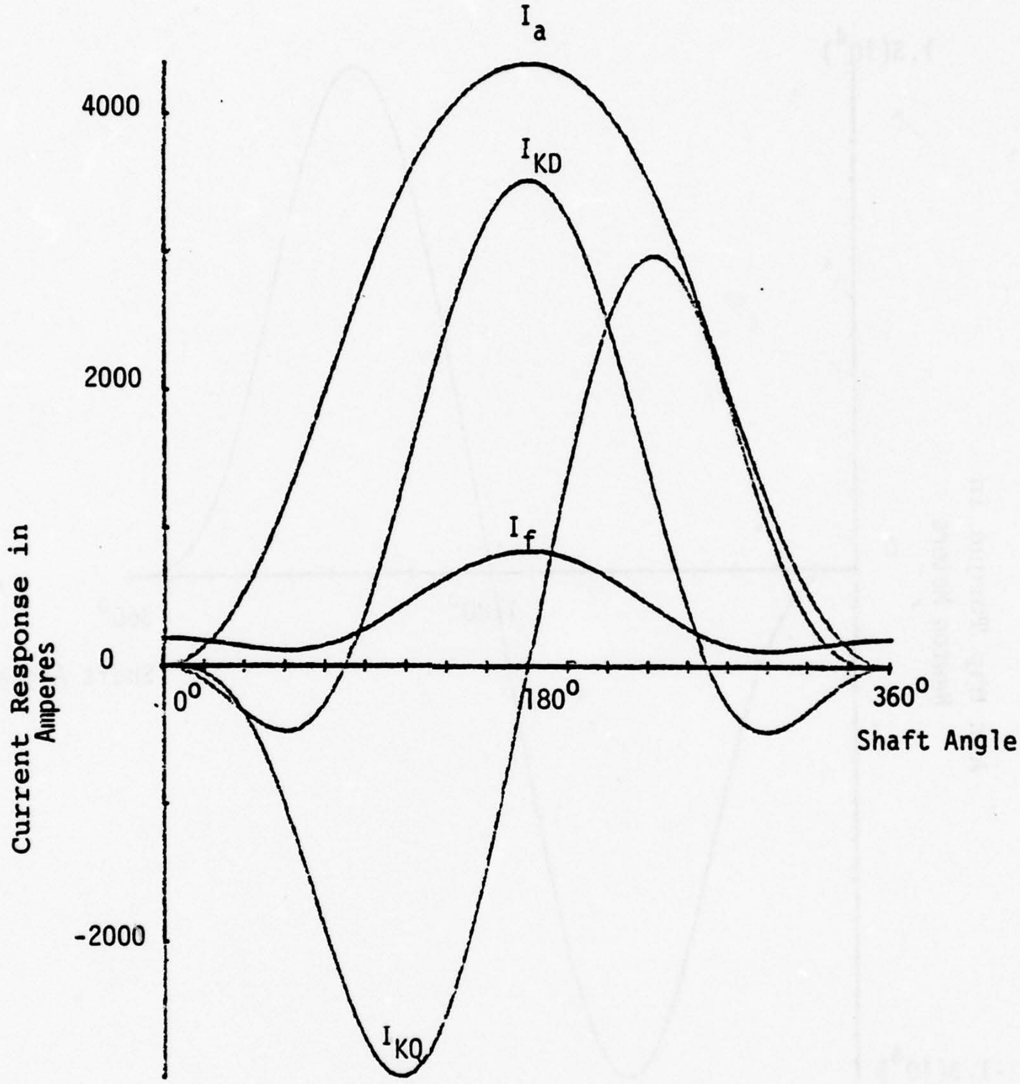


Figure 5. Phase A to ground current response

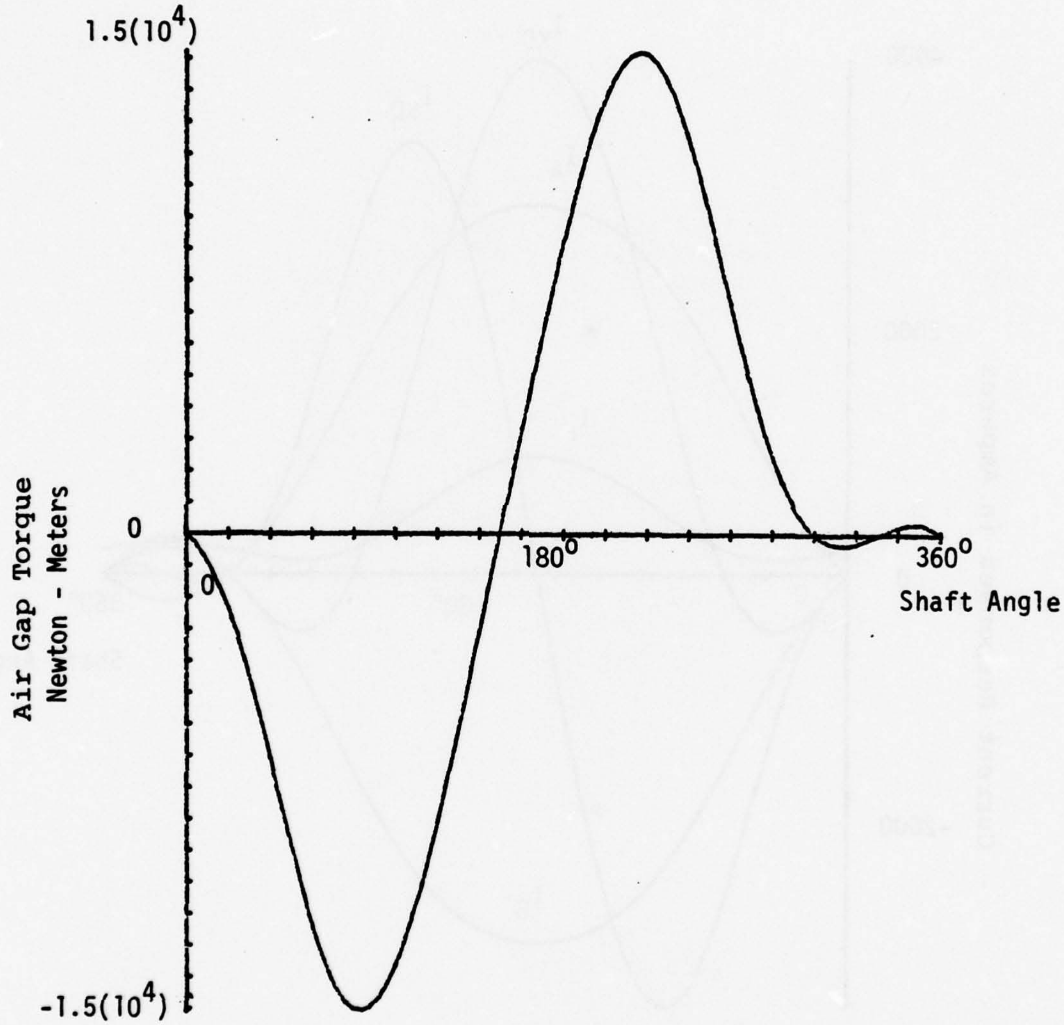


Figure 6. Phase A to Phase B fault torque response

I_A = phase current
I_F

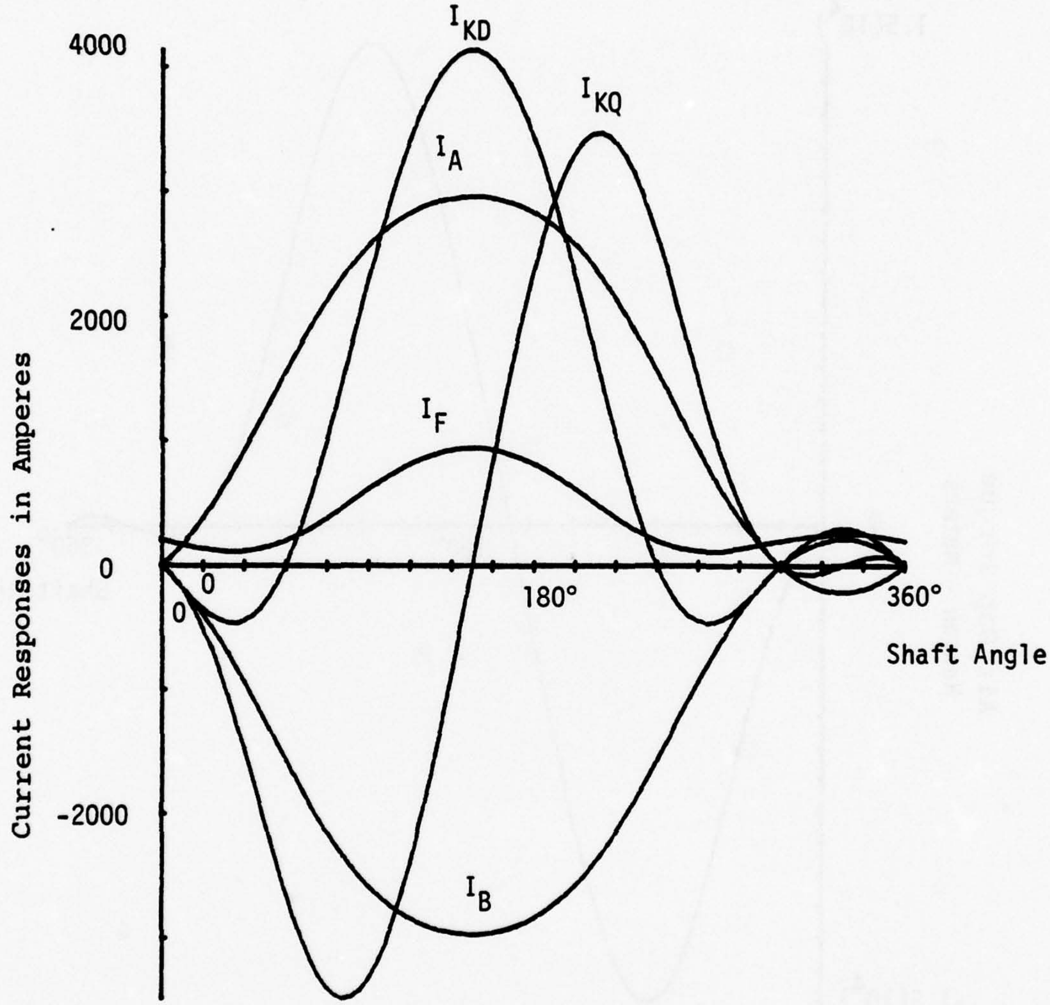


Figure 7. Phase A to Phase B short; current response

FILTER WEIGHT AND VOLUME

The basic circuit configuration for the rectifier/filter combination is shown in Figure 8. There are several choices for the circuit configuration of the filter. The specific configuration is usually determined by cost, availability of components, and desired efficiency. Because of the intended application, the filter is a point design with an optimum weight and volume requirement. The most effective filter circuits involve inductors (L) and capacitors (C). Because multiple L-C section filters do virtually the same job as a single L-C section (the choice is one of available component sizes), the simple L-C filter of Figure 8 is chosen for the base design.

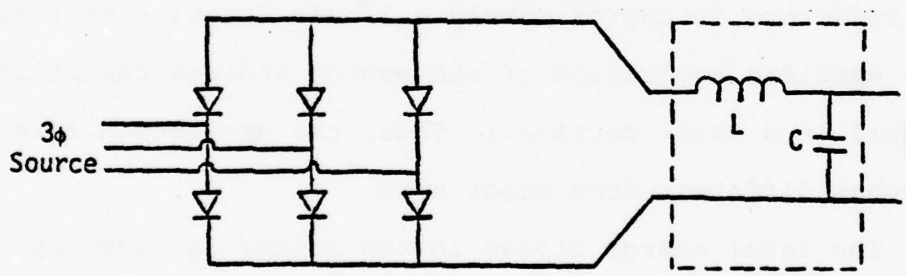


Figure 8. Basic filter circuit

The optimum inductor configuration is generally conceded to be the so-called Brook's coil. However, this design generates a considerable field external to the coil. Shielding

is necessary to comply with Mil Std. 461A. A toroidal coil design, on the other hand, does not produce excessive external fields and requires little or no shielding. The choice of air core or an iron-gap configuration is not clear. Air core designs are investigated here.

The degree of filtering action or ripple factor is proportional to the inverse of $\sqrt{LC}$. Thus, many choices exist for the component values of inductance and capacitance for the same ripple factor. To determine what the optimum choices are, a formulation of filter weight in terms of component energy storage capability can be made (ref. 1). Basically, this formulation expresses filter weight as a function which relates weight to energy, i.e. pounds/joule (lb./J). For capacitors, this factor is a constant with volume but for inductors is approximately a linear function of dimension. (The specific evaluation of the energy storage capability is covered in a later section.) Thus, the evaluation here is somewhat different from prior work.

The total energy stored in the filter is given by the equation

$$W = 1/2\ LI_o^2 + 1/2\ CE_o^2 \qquad (2)$$

where

W = energy in joules (j)
L = inductance in henries (h)
C = capacitance in farads (f)
I_o = D.C. load current in amperes (amps)
E_o = D.C. load voltage in volts

The factors K_C and K_L are defined as:

K_C = lb/J *for* the capacitor

K_L = lb $h^{1/3}$/Joule for the inductor.

Now, the weight of the filter can be expressed as:

$$\text{Weight} = 1/2\ L^{2/3} I_o^2 K_L + 1/2\ C E_o^2\ K_C \qquad (3)$$

The values of inductance and capacitance are related by the expression

$$\gamma = \frac{0.0402}{\omega^2 LC} \qquad (4)$$

were γ = ripple factor, the ratio of ripple voltage to E_o

ω = ripple frequency.

Rewriting this expression for C,

$$C = \frac{0.0402}{\omega^2 \gamma L} \qquad (5)$$

The filter weight can now be expressed as

$$\text{Weight} = 1/2 L^{2/3} I_o^2 K_L + 1/2\ \left(\frac{0.0402}{\omega^2 \gamma L}\right) E_o^2 K_C \qquad (6)$$

By taking the derivative of this expression with respect to L and equating the result to zero, an expression for L is found:

$$L = \left[\left(\frac{0.0603}{\omega^2 \gamma} \frac{K_C}{K_L}\right) (R_o)^2\right]^{3/5} \qquad (7)$$

where $R_o = \dfrac{E_o}{I_o}$.

The corresponding value of capacitance is

$$C = \frac{0.217}{\omega^{4/5}\ \gamma^{2/5}\ R_o^{6/5}} \frac{(K_L)^{3/5}}{K_C} \qquad (8)$$

An evaluation of filter weight using an air core toroid can now be made.

The equation for the inductance of a single layer, air core toroidal inductor is given by (ref. 2):

$$L = 4\pi 10^{-7}\ N^2 \left[R - \sqrt{R^2 - a^2} \right] (h) \qquad (9)$$

where R and a (meters) are defined in Figure 9 and N is the number of turns. The circumference of the inner portion of the toroid is given by $2(R-a)\pi$. Thus, the number of turns that can be wound in a single layer is given by

$$N = \frac{2(R-a)\pi}{d}$$

where d is the wire diameter including insulation. The expression for L is now

$$L = \frac{16\pi^3\ 10^{-7}\ (R-a)^2}{d^2} \left(R - \sqrt{R^2 - a^2} \right) \qquad (10)$$

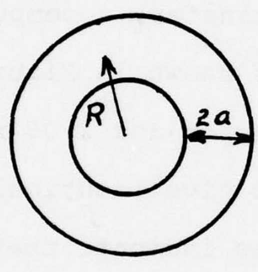
Taking the derivative of the inductance of equation 10 with respect to a and equating the result to zero, it is found that maximum inductance occurs for

$$a = 0.5205\ R$$

An estimate of the inductance of a two layer winding can be made by assuming the coefficient of coupling between the two layers is 0.95. The total inductance is then given by the equation

OPTIMUM TOROID

$$L = \frac{16\pi^3(10^{-7})(R-a)^2}{d^2}\left[R-\sqrt{R^2-a^2}\right]$$



$a = .5205\,R$

GREATEST INDUCTANCE FOR GIVEN AMOUNT OF WIRE

Figure 9. Torous for coil form

$$L_t = L_1 + L_2 + 1.90\sqrt{L_1 L_2} \qquad (11)$$

where L_1 and L_2 are the inductances of the individual layers. It is assumed that if the torus dimensions are optimum for a single layer winding, it will be approximately so for a two layer winding.

To evaluate the inductance, a rational basis for choosing the wire size must be established. Allowable temperature rise is used for this analysis. A 200°C rise from an initial temperature is assumed to be reasonable. Utilizing the heat equation for no heat transfer, a computer solution for various initial temperatures is shown in Figure 10. The current density is 3.13×10^7 A/m^2 for copper and 2.06×10^7 A/m^2 for aluminum. These current densities give identical temperatures versus time responses. These curves indicate that the stated current densities are acceptable for 30 seconds of operation without cooling.

Using the above current densities to determine wire size and preceding equations for inductance, a series of inductors can be designed using the derived optimum configuration. Shown in Figure 11 are curves for inductance versus R for aluminum and copper wire with insulation for 20 kV operation. An insulation stress of 200 V/mil was used in those calculations.

If the inductors are wound on a 1/16inch epoxy-glass coil form (density of 2) and the insulation has a density of 1, an

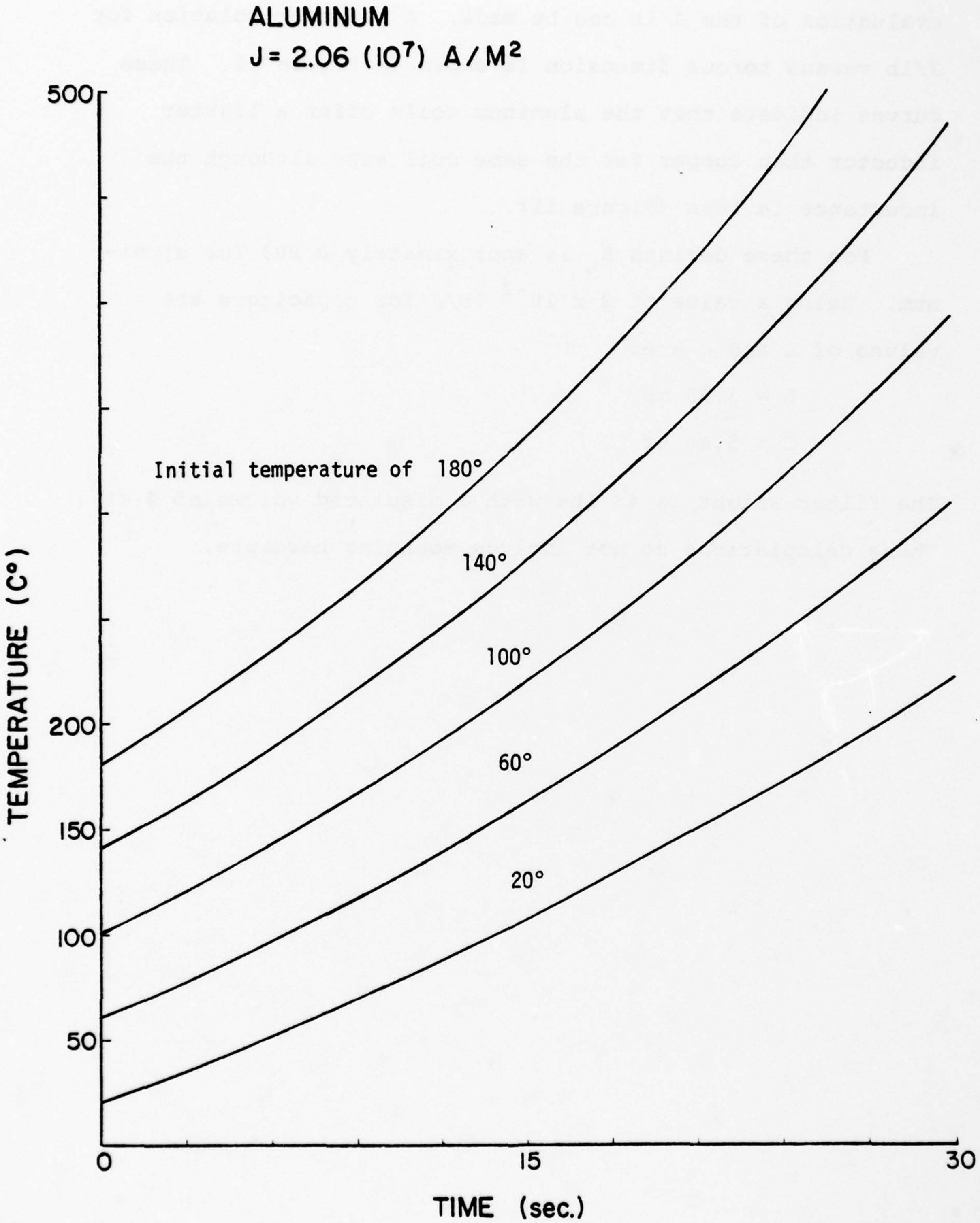


Figure 10. Temperature versus time

evaluation of the J/lb can be made. A computer solution for J/lb versus torous dimension is shown in Figure 12. These curves indicate that the aluminum coils offer a lighter inductor than copper for the same coil size although the inductance is less (Figure 11).

For these designs K_L is approximately 0.203 for aluminum. Using a value of 2×10^{-2} lb/J for capacitors the values of L and C are:

$$L = 3.55 \text{ mh}$$

$$C = 5.46 \ \mu f$$

The filter weight is 43 lbs with a displaced volume of 4 ft^3. These calculations do not include mounting hardware.

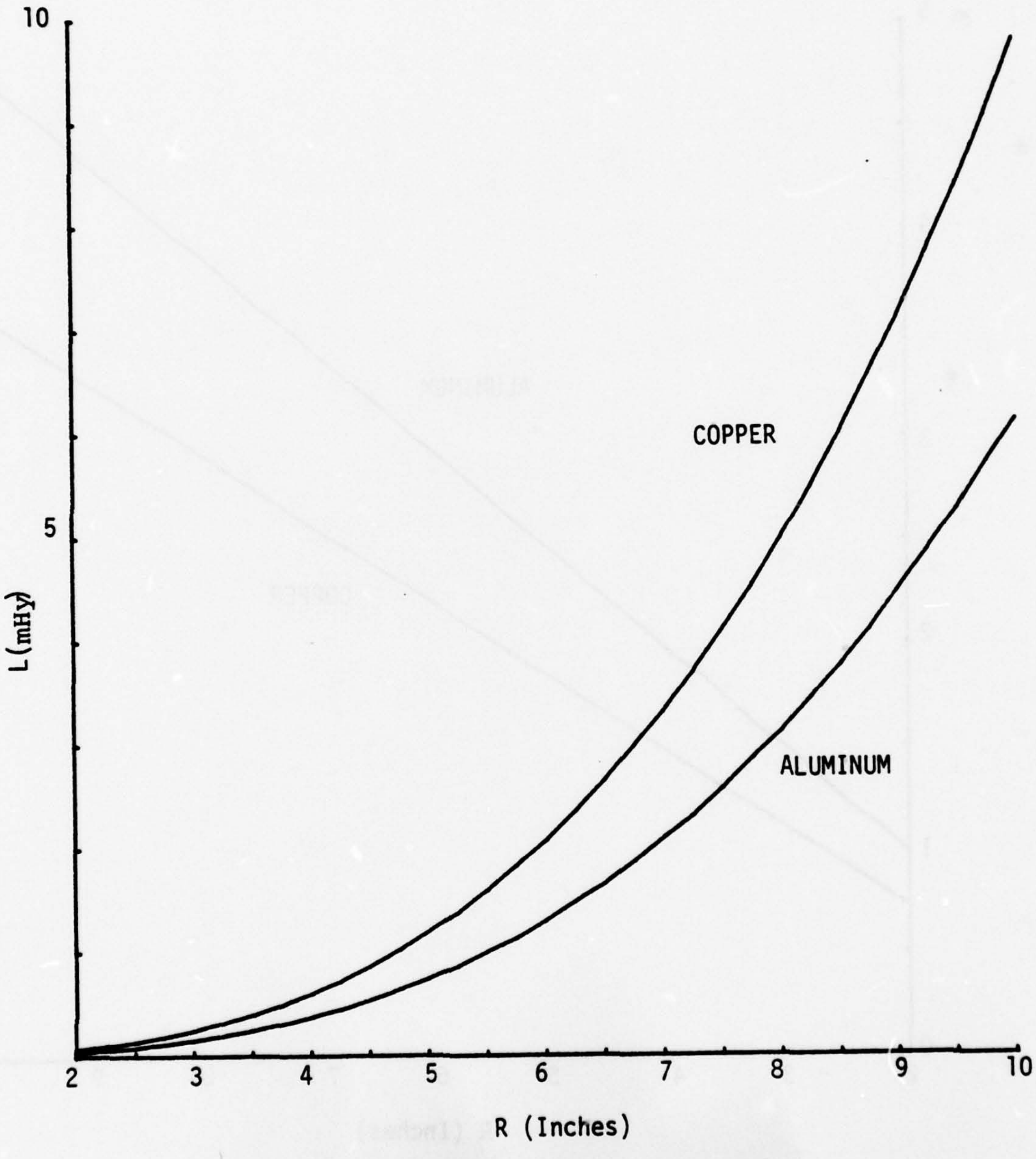


Figure 11. Inductance versus toroid dimension

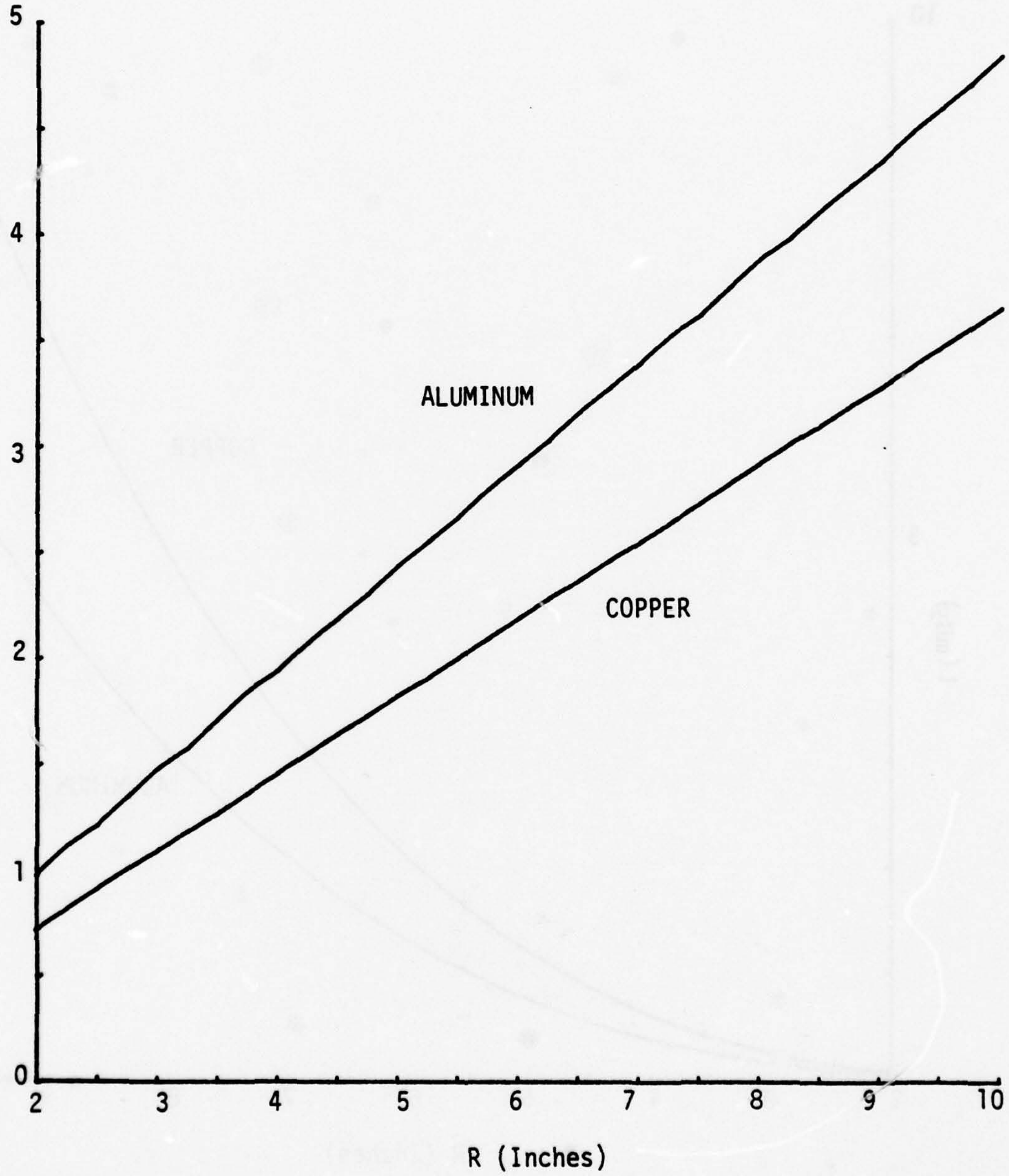


Figure 12. Energy per unit weight of inductor for I_o = 235 amp.

PACKAGING

Packaging of power conditioning equipment is a complex problem with many factors to consider simultaneously. For instance, shock loads, crash loads, and heat control as well as insulation for the high voltage are among the more important considerations. Clearly, a detailed design is of little value without a specific application (pod design, etc.). However, certain generalizations can be made. Because the ultimate utilization will be airborne, the container will certainly be pressurized. This constraint along with the lightweight requirement suggests the use of a cylindrical container that can be pressurized with an insulating gas. A conceptual layout of such a container with the major components is illustrated in Figure 13.

The SCR stacks form the rectifying bridge. The electrical connections to the bridge and the location of the SCR controls are indicated in Figure 14. Note the simplicity of the A.C. connection to a three phase source. By utilizing a toroidal inductor and a capacitor concentric to the inductor, a small filter volume can be achieved (see Figure 15).

The location of the crowbar components is shown in Figure 13. Of course a very detailed analysis is required to establish compatability of the entire layout. It is possible, for instance, to attach the power conditioning container directly

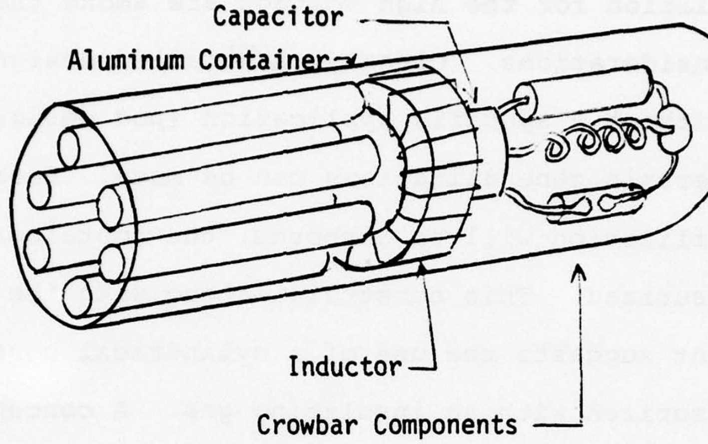


Figure 13. Conceptual layout of power conditioning equipment

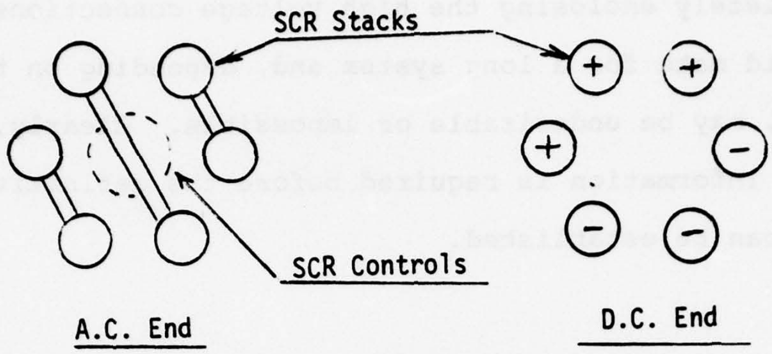


Figure 14. End views of SCR stacks showing electrical connections

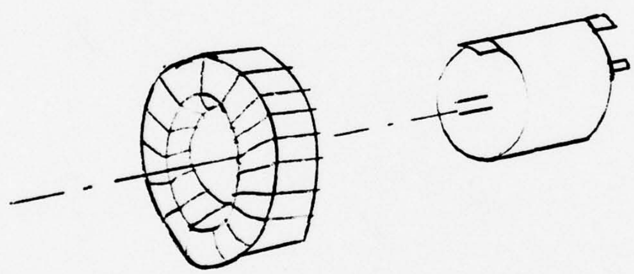


Figure 15. Expanded view of filter components

to the alternator housing, thus eliminating one end bell and completely enclosing the high voltage connections. This would make for a long system and, depending on the airplane, may be undesirable or impossible. Clearly, more detailed information is required before the definitive designs can be established.

CROWBAR PROTECTION OF LOAD

The objective of the crowbar is to reduce the amount of energy deposited in a load fault and thus reduce load damage to an acceptable level. Because of the energy stored in the filter and the follow-through capability of the APU, the available fault energy must be diverted from the load. This can be accomplished in part by the circuit shown in Figure 16. The inductor L_p is made small compared to the filter inductor. In the event of a load fault, a substantial portion of the filter capacitor voltage will appear across L_p. The voltage appearing at the trigger electrode to the spark gap will then be sufficient to achieve firing of the spark gap. The available fault energy is then diverted from the load, at least in principle. The inductor, L_p, then serves the dual purpose of triggering the crowbar and limiting the initial current in the load fault.

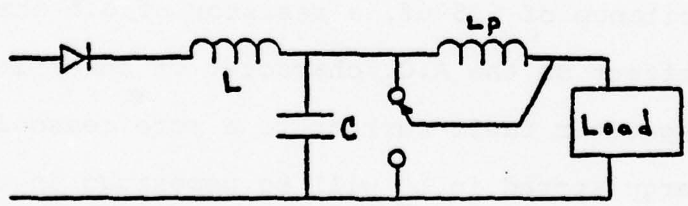


Figure 16. Simple load fault protection circuit

The circuit of Figure 16 must be modified because of the fragile nature of the load. First there is no physical resistor to absorb the energy in the filter capacitor. The best location for a resistor to perform this function is in series with the filter capacitor as shown in Figure 17.

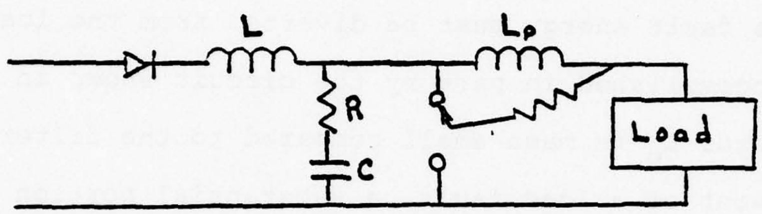


Figure 17. Improved load fault protection

This location allows the resistor, R, to absorb the capacitor energy under fault conditions and does not interfere with D.C. load current. Placement of this resistor in series with the spark gap would raise the voltage applied to the load fault and would diminish the effectiveness of the crowbar. For a filter capacitance of 5.5 μf, a resistor of 0.5 ohm will have negligible effect on the A.C. character of the filter and will limit the capacitor fault current to a more reasonable value.

The energy stored in L_p will be deposited in the load arc, and the voltage drop across the spark gap may be too large to allow the load arc to extinguish. The energy stored in the connecting cables and stray capacitances will also be

deposited in the load. To further reduce energy in a load fault a more complex circuit is required.

A few simple relationships are useful for design considerations. The circuit shown in Figure 18 roughly approximates the conditions existing after the crowbar has fired if the spark gap voltage is neglected for the moment.

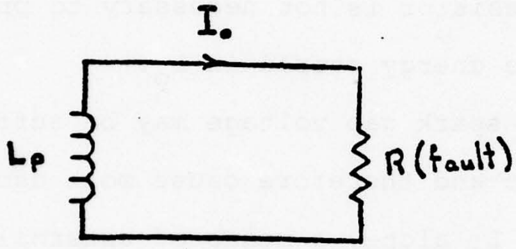


Figure 18. Model of circuit after crowbar fires (without spark gap voltage)

The decay in the fault current is given by the equation

$$i(t) = I(0)\varepsilon^{-t/\tau} \qquad (12)$$

where the time constant, τ, is L_p/R (R is assumed to be linear for this analysis). By integrating i(t) with respect to time an expression for the charge transport through the load is obtained:

$$Q = \int_0^{\infty} i(t)dt = \frac{L_p}{R} I(0) \qquad (13)$$

This expression is useful for estimating the damage caused by the energy stored in the inductor, L_p. Well designed spark gaps have erosion rates of between 10^{-4} and 10^{-5} gm/c

of charge transferred. If 10^{-3} gm is the maximum allowable amount of material transferred during a fault, then between 10 and 100 coulombs is the maximum allowable charge transfer. Using 50 μh and 235 A for L_p and I(0) respectively, the fault resistance is determined to be between 1.2 and 12 milliohms. This value is very low and the resistance of the leads and L_p will exceed this value. Thus, from an erosion point of view, an extra resistor is not necessary to prevent excessive damage due to the energy stored in L_p.

Because the spark gap voltage may be sufficient to maintain the load arc and therefore cause more damage than the energy stored in L_p alone, a means of externally extinquishing the load arc may be useful. The load arc is extinguished when a current zero occurs. This may be accomplished by forcing a current zero by additional circuit elements. The circuit shown in Figure 19 can automatically achieve a zero current in the load by discharging the capacitor C_p into the load. The capacitor discharge current is in a direction opposite to that of the fault current. Of course, the energy deposited in the fault by the capacitor must be smaller than would otherwise be the case. Extinguishing an arc in this manner is referred to as forced commutation.

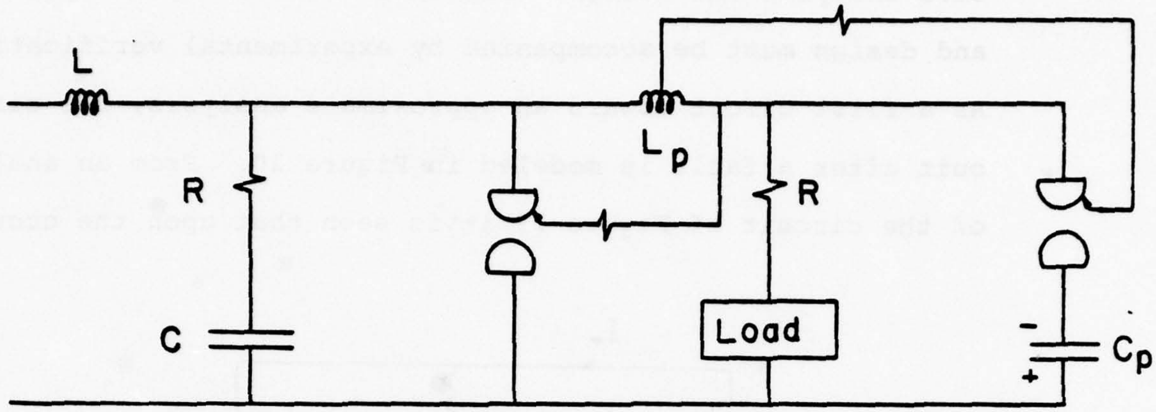


Figure 19. Improved load fault protection circuit

The exact character of arc impedance is very non-linear and hard to express in analytical terms for definitive analysis and design. For this reason, any analysis and design must be accompanied by experimental verification. As a first effort toward an approximate analysis, the circuit after a fault is modeled in Figure 20. From an analysis of the circuit of Figure 19 it is seen that upon the occurrence

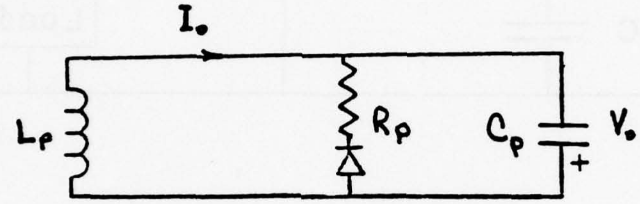


Figure 20. Model of improved load fault circuit after crowbar is fired

of a load fault both spark gaps are triggered, and the circuit is approximated by that of Figure 20. The diode in the load circuit in series with the resistance R_p is intended to simulate the arc character near zero current. At low arc current it is postulated that the arc resistance R_p increases rapidly and near the zero current becomes large enough so that the parallel RLC circuit of Figure 20 becomes underdamped. This result is easily seen by solving the circuit differential equation. Thus, L_p and C_p form a resonant circuit with a sinusoidal

response as the capacitance C_p discharges. If this is even approximately the case, computer analysis indicates that the fault voltage will be maintained at a low value for sufficient time for the load arc to extinguish. Figures 21 and 22 are computer solutions of the circuit shown in Figure 20. The circuit values shown in Figure 21 and 22 are not completely realistic and much more analysis and experimental verifications are required.

First indications are, however, that a simple and reliable method of fault protection can be derived from the idea of forced commutation of a load arc. Many problems are apparent in this simple analysis. However, an alternative method of fault management is possible with very fast recovery.

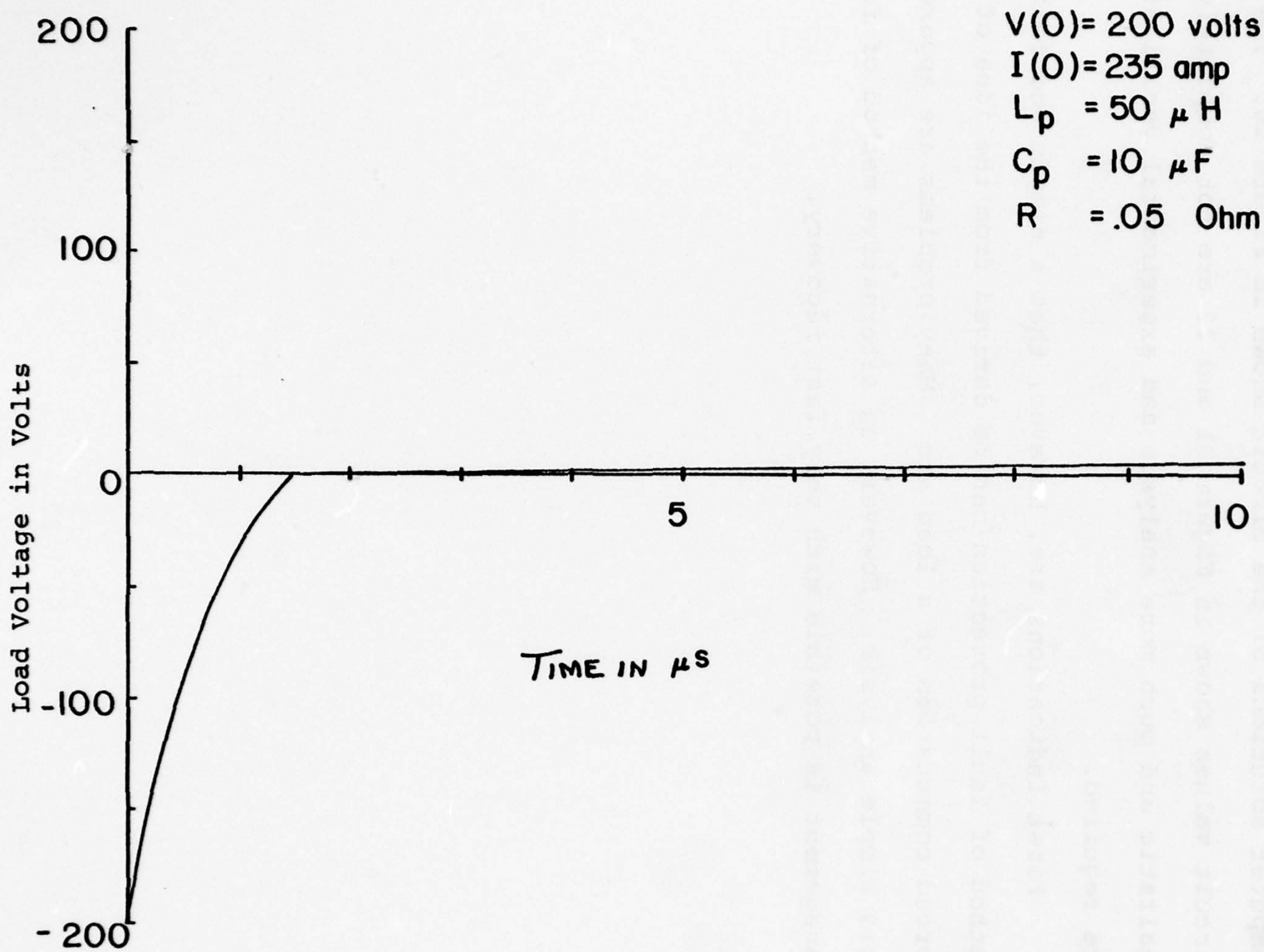


Figure 21. Fault voltage versus time at R = 0.05 ohm

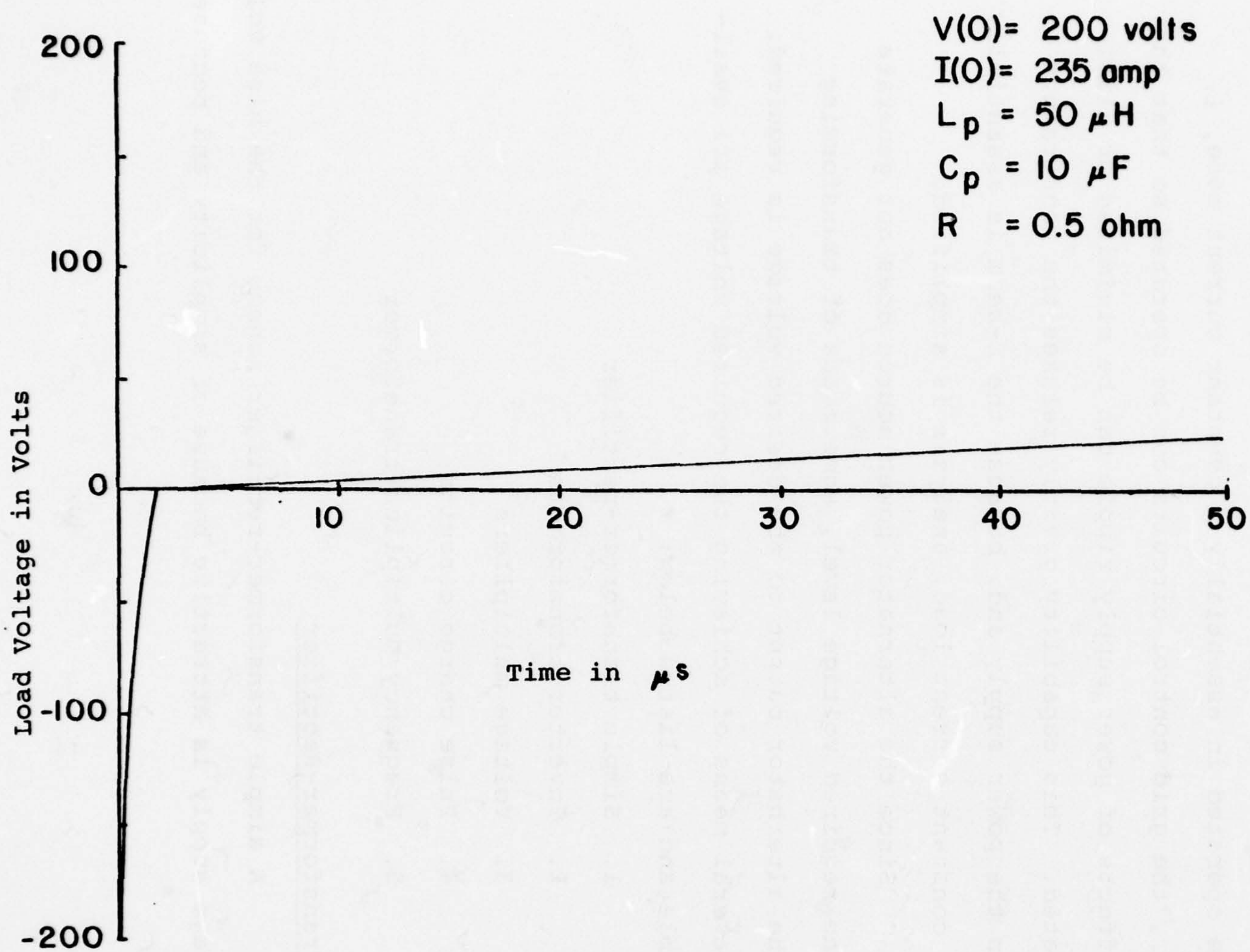


Figure 22. Fault voltage versus time at \$ = 0.5 ohm

E-BEAM SUPPLY

The power supply for the E-beam gun has to meet, at least in principle, fewer constraints than does the main load. The principal function of the E-beam is to control load impedance. Since the gun has a control grid, it can be operated in essentially a constant current mode, i. e., the grid control circuits can be operated so that the effects of power supply ripple can be minimized or eliminated. This capability greatly relaxes the constraints on the power supply and, because the E-beam is essentially a constant current load, analysis is simplified.

Since the alternator power source does not generate the required voltage level, some means of transforming the alternator output to the desired voltage is required. Several means of achieving the required voltage are available and are listed below:

1. Simple transformer-rectifier
2. Inverter-transformers
3. Voltage multipliers
4. Pulse charge circuits
5. Frequency multiplier-transformer

Transformer-Rectifier

A simple transformer-rectifier scheme for the high voltage supply is attractive because of simplicity and because

this is the classical method for generating high voltage D.C. at significant power levels. The simplest scheme is a single phase transformer with a full or half wave rectifier. Polyphase rectifying schemes offer a much reduced output ripple and a better loading of the alternator source than a single phase scheme. A first order comparison of weight for a polyphase (3ϕ) and single phase transformer/rectifier scheme can be easily made by noting that transformer weight is proportional to the kVA rating raised to the 3/4 power (ref. 3). For a power output of 200 kW three 67 kVA transformers are required for a three phase scheme. (A single 3ϕ transformer is smaller than three 1ϕ transformers but usually has 5 to 6 percent phase unbalance or does not package well.) Thus, the ratio of transformer weight is given by

$$\frac{W_{3\phi}}{W_{1\phi}} = \frac{3(67000)^{3/4}}{(200,000)^{3/4}} = 1.32 \qquad (14)$$

The conclusion is that the transformers (and essentially the power supply for the output current requirement) for the three phase rectifier scheme is approximately 30 percent heavier than for a single phase scheme.

Projections made from classically designed transformers give rather large weights for the transformer(s). However, because the duty cycle is very limited, a heat sink design with little or no cooling will result in more reasonable

designs. Calculations for transformer weight in this power range utilizing 1.8 Tesla, 26×10^6 A/m^2, and reasonable winding factors indicate that 0.1 lb/kW is reasonable for the required iron and copper in this power range. Packaging will increase the weight so that a factor of 0.15 lb/kW is used for evaluation in this analysis. Thus, the weight for 200 kW is approximately 30 lb. Because of the small weight, a three phase scheme may be reasonable especially if low ripple is required.

A reasonable weight estimate for a single phase transformer-rectifier at the required voltage and current levels seems well within the 150 lb total weight if SF_6 is used as an insulating medium. Because the load can be compensated for ripple, filter requirements are minimal and a simple capacitor filter should suffice. A capacitance of 0.2 μf would give a regulation of 10 percent for a 1/2 wave rectifier. Of course, the effect of a transformer/rectifier load on the alternator and main load must be determined before a completed design can be established.

Inverter-Transformer

Designs and estimates of weights of inverters have been adequately covered elsewhere. This discussion is limited to the pros and cons of an inverter in the present application. Because inverters achieve light weight due to high

frequency operation, filter requirements are much reduced. Thus, the output filter capacitor is much smaller and consequently stores much less energy than would otherwise be the case (transformer/rectifier scheme). Because small high voltage devices are likely to arc, reduced arc energy is an obvious system advantage. Also, because the inverter uses switches at high frequency, the possibility exists of rapidly interrupting power flow into an arc so that little or no protection equipment may be needed with the use of an inverter in a high voltage power supply.

Another advantage that some inverter designs offer is the possibility of output voltage regulation that is reasonably independent of the supply voltage. As the main supply is likely to be regulated based on the main load requirements, the advantage of independent voltage control for the auxiliary load may become an important consideration. An inverter power supply scheme would serve to decouple the main load from the auxiliary load, thus assisting in preventing load interactions leading to unforeseen system instabilities.

Voltage Multiplier

An alternative method of obtaining high voltage is the Crockcroft-Walton voltage multiplier circuit. The three stage circuit configuration shown in Figure 23 will result

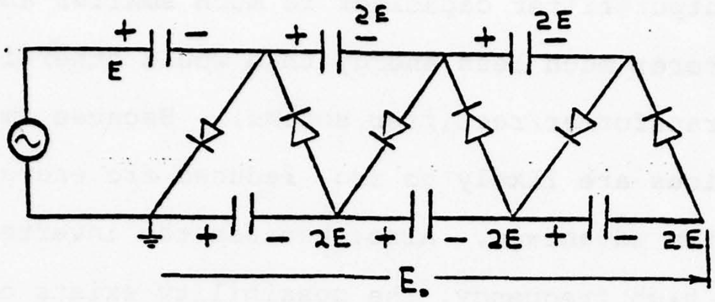


Figure 23. Crockcroft-Walton Voltage Multiplier

in an ideal charge voltage of 2E on each of the capacitors except the first (C_1) where E is the peak A.C. voltage of the source. If the output voltage is taken as indicated in Figure 23, six times the peak source voltage is available. However, the output voltage will have a ripple at the source frequency characteristic of a 1/2 wave rectifier.

Because of the simplicity of the Crockcroft-Walton circuit, it is interesting to consider how it might be used in the anticipated application. To directly charge this circuit from the alternator as shown in Figure 23 is impractical because a ground is required at the alternator. Such a ground would short the rectifiers for the main load. To circumvent this difficulty, consider the circuit shown in Figure 24. Note that the ground has been placed so that no D.C. path exists between the alternator and ground. The voltage on

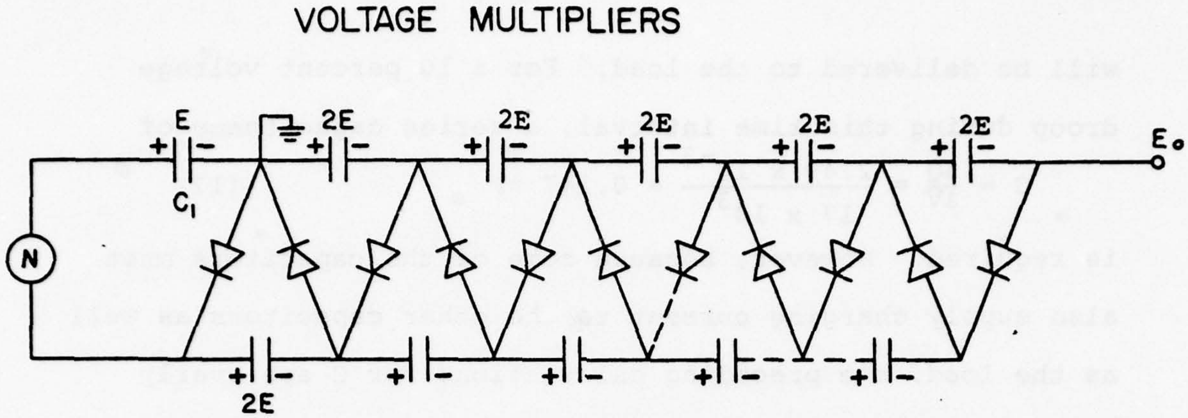


Figure 24. A multistage Crockcroft-Walton circuit for use with an ungrounded source

capacitor C_1 places a D.C. stress on the alternator insulation, however, and this effect must be evaluated by the alternator designer. If the voltage E shown in Figure 24 is 17 kV, then the output voltage, E_o, will be 170 kV.

An estimate of the performance of this circuit can be made without performing a complete detailed analysis. A peak-to-peak voltage ripple of 10 percent and a constant current characteristic of the E-beam load are assumed.

For a 1/2-wave rectifier, the time between charging pulses is given by

$$T = \frac{1}{f} = \frac{1}{562} = 0.00178\,\mathrm{s} \qquad (15)$$

For a constant load current of 1.4 amperes a total charge of

$$Q = (1.4)\ (0.00178) = 2.49 \times 10^{-3}\ \mathrm{C} \qquad (16)$$

will be delivered to the load. For a 10 percent voltage droop during this time interval, a series capacitance of

$$C = \frac{\Delta Q}{\Delta V} = \frac{2.49 \times 10^{-3}}{17 \times 10^{3}} = 0.147\ \mu F \qquad (17)$$

is required. However, because some of the capacitors must also supply charging current to the other capacitors as well as the load, the preceding calculations for C are overly simplified. The average output voltage of a Crockcroft-Walton voltage multiplier is given by (ref. 4) the equation

$$V_o = 2mV - \frac{I_1}{fC}(2/3\ m^3 + 1/2\ m^2 - 1/6\ m) \qquad (18)$$

for all the capacitors equal in value where m is the number of stages. Solving for C with the preceding values of ΔV, I_1, and f and with m = 5,

$$C = 13.9\ \mu f \qquad (19)$$

A quick evaluation of weight using 100 J/lb reveals a weight estimate of approximately 850 lbs for the required voltage level. This weight is clearly excessive and voltage multiplier circuits will not be considered further.

<u>Pulse Charge Circuits</u>

The use of pulse charge schemes offers the possibility of very lightweight components in a high voltage power supply. The trade-off is, as in the case of inverters, in the transformer. The method presented here utilizes a circuit

commonly used in large electromagnetic pulse (EMP) generators to transfer charge from a slow pulser to a fast pulser. Figure 25 is a simplified circuit of this scheme. Usually the capacitors C_1 and C_2 are equal and 100 percent charge transfer is possible between C_1 and C_2. As it is desirable from a D.C. supply point of view to have complete charge transfer from a small capacitor (C_1) to a large filter capacitor, conditions necessary for this to occur can be found by requiring the energy to be zero in C_1 at the end of the charging cycle. Calculations indicate that the charging voltage as well as the initial voltage on C_2 (load voltage) is related to the values of C_1 and C_2. Obviously, if $E_1(0) < E_2(0)$, no charge transfer takes place.

By using another circuit to resonantly charge C_1 to a low voltage, C_1 can then be discharged through a step-up pulse transformer to achieve charge transfer. If a 20 to 30 kHz repetition rate is used, then most of the circuit elements can be made very small. The resonant inductance indicated in Figure 25 can in principle be realized using the leakage inductance of the pulse transformer. A circuit of this pulse charge scheme is shown in Figure 26. The capacitor C_2 is chosen based on the allowable load voltage droop during the intercharge period. C_1 is determined by

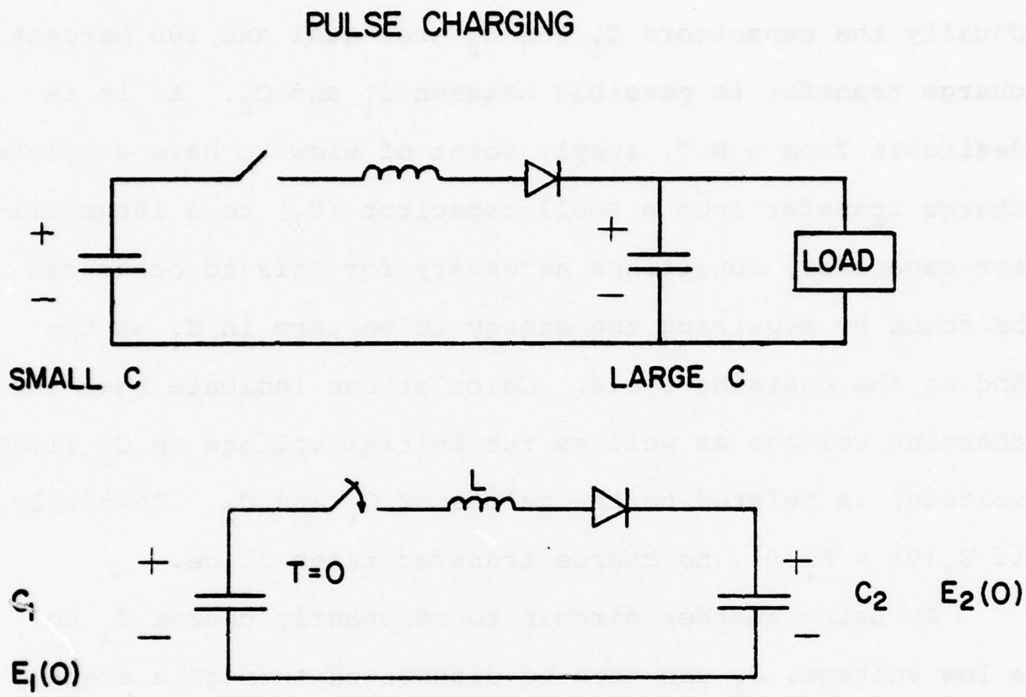


Figure 25. Pulse charging networks

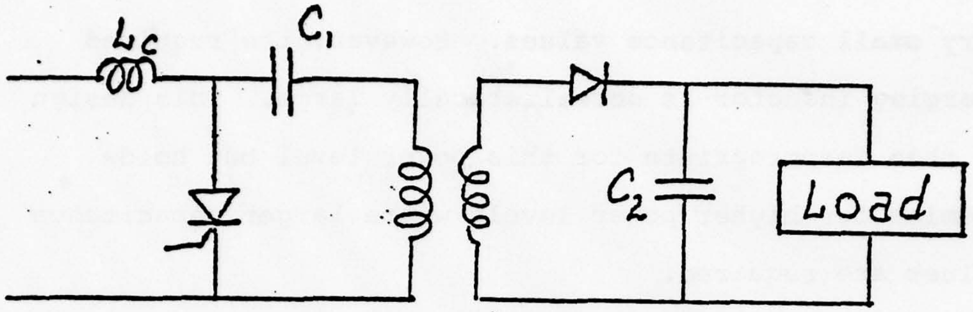


Figure 26. Pulse charge, high voltage supply

the supply voltage and the required energy flow. The turns ratio of the pulse transformer in Figure 25 is determined by the charge transfer. The charge transfer period can be made 10 to 20 percent of the intercharge period.

Under load conditions the charge transfer characteristics are modified from the ideal case. If the load is represented as a constant current drain, the charge transfer voltage is given by the expression

$$e_2(t) = E_2(0) + \frac{E_1'C_1'}{C_2}\left[1-\cos\frac{t}{\sqrt{LC_1}}\right] + \frac{1}{C_2}\left[\frac{(I_o C_1')}{C_2} - I_o\right] t$$

$$- \frac{\sqrt{LC}'C'}{C_2^{2}} \; I_o \sin\frac{t}{\sqrt{LC'}} \qquad (20)$$

where

$$C' = \frac{C_1 C_2}{C_1 + C_2}$$

A power supply design using this scheme and a 30 kHz repetition for the required power level results in very small capacitance values. However, the required charging inductor is unrealistically large. This design is then inappropriate for this power level but holds promise for higher power levels where larger capacitance values are required.

It is seen that choices for the power supply schemes reduce to the transformer/rectifier or the inverter scheme. The transformer/rectifier scheme is by far the simplest and most cost effective. However, inverters operating at high frequencies offer better overall system advantages. For instance, a smaller filter is required reducing the energy available for arc damage and, because of the high frequency, more rapid fault interruption is possible. Also, some inverter designs offer an easy method of voltage regulation if this becomes necessary. Thus, the final choice of a power supply scheme should be based on overall system requirements.

Frequency Multiplier-Transformer

For some pulse power applications, electrical power is generated at a convenient voltage and frequency; the voltage is subsequently transformed to a higher level. Since the

size and weight of the transformer is a function of the frequency, it is sometimes desirable to change the frequency of the generated voltage ahead of the transformer. Solid state converters and inverters may be used for the frequency changer. In this arrangement, the input and output frequencies may be single or polyphase, and the number of phases for the output frequency need not equal the number of phases for the input.

An alternative and similar procedure which may permit a simpler switching arrangement is to omit the D.C. portion of the above and use a static frequency multiplier. Most high power level frequency converters in the past have used saturable magnetic core devices, and this is a possibility for this application. However, semiconductor switches can also be used, and they may prove to be more suitable with regard to weight, cost, and efficiency. The saturable magnetic devices may be more rugged and the same cores may be utilized for the step-up transformer. Regardless of which switching devices are used, the frequency multiplier can be designed with different numbers of phases and frequency ratios.

As an illustration, a frequency tripler is shown in Figure 27, with a three phase input and single phase output.

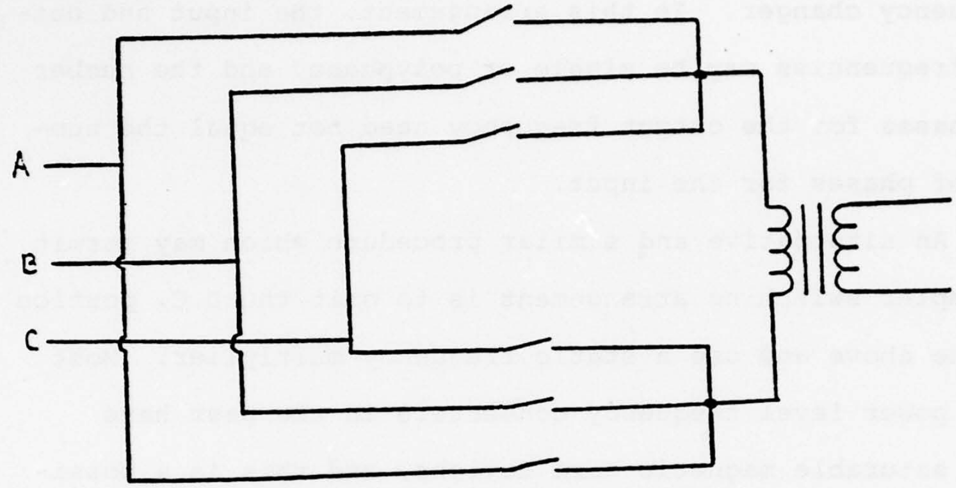


Figure 27. A frequency tripler

COMMON POINT GROUNDING

For most electric circuits, the exact nature of the grounding scheme is not of major significance. However, at high frequencies where circuit dimensions become comparable to electrical wavelengths and at high power levels with accompanying high currents, the grounding scheme becomes important to insure proper system operation. Proper grounding helps prevent interference between otherwise independent circuits and prevent damage to sensitive components. The high power level is of more importance for the present application and is addressed in some detail in succeeding paragraphs.

Ordinary circuit diagrams do not illustrate grounding schemes in sufficient detail to be meaningful. The physical realization of the circuit and the electrical schematic usually are not recognizable to anyone but the designer and technicians. To the designer of low power systems, certain conditions assumed in that design may be translated to work associated with high power design. These assumptions are primarily of magnitude and not of principle. A ground conductor with 10 milliohm resistance is adequate for logic design; but if this conductor is carrying 40 kA, then a 400 V potential will exist across the conductor. Such voltages will damage integrated circuits.

As an example to illustrate the effects of the grounding scheme, consider the circuit layout shown in Figure 28a.

Major components are shown grounded to an aluminum ground plane. For simplicity, the ground plane is assumed to be infinite in extent. By solving LaPlace's equation, the current distribution and equipotential lines can be obtained. For a current injected into point B and removed at point A, the current distribution and equipotential lines are shown in Figure 28b. If the current is large, then obviously points C and F are not at ground or zero potential. A potential or common mode voltage will exist between the control box (F) and the amplifier (C), which may cause damage or erroneous system operation. Even with the assumption of an infinite ground plane, a 50 Hz, 40 kA current injected and removed at points separated by 2 m, the potential existing between the points is 7.5 V. Because of the skin effect and finite conductivity of practical conductors (in this case aluminum), even an infinite ground plane cannot be maintained at zero potential. The designer must keep in mind that fault conditions can occur so that even if normal operation is satisfactory, a one time fault may cause vital circuitry to malfunction or fail.

The preceding example, though impractically large, serves to illustrate the effect of common ground return conductors. Practical circuits cannot be so simply visualized, and certainly the effect of ground currents includes using digital signals large compared to the common mode voltage and differential input amplifiers for analog signals. These

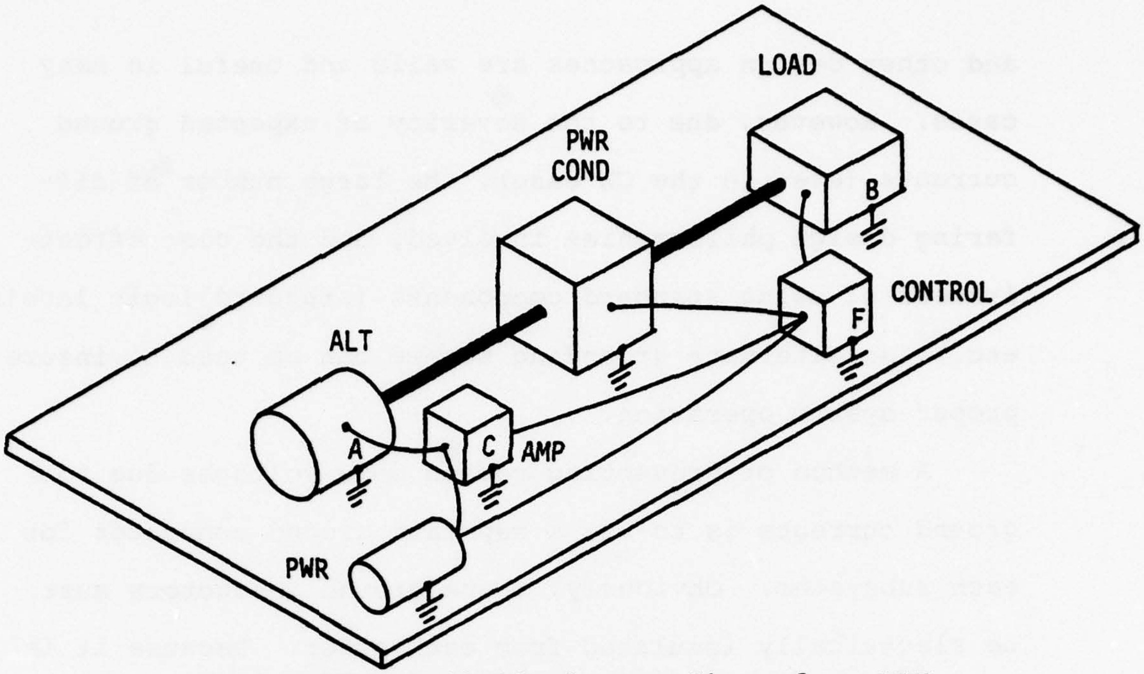


Figure 28a. Individual grounding of an APU

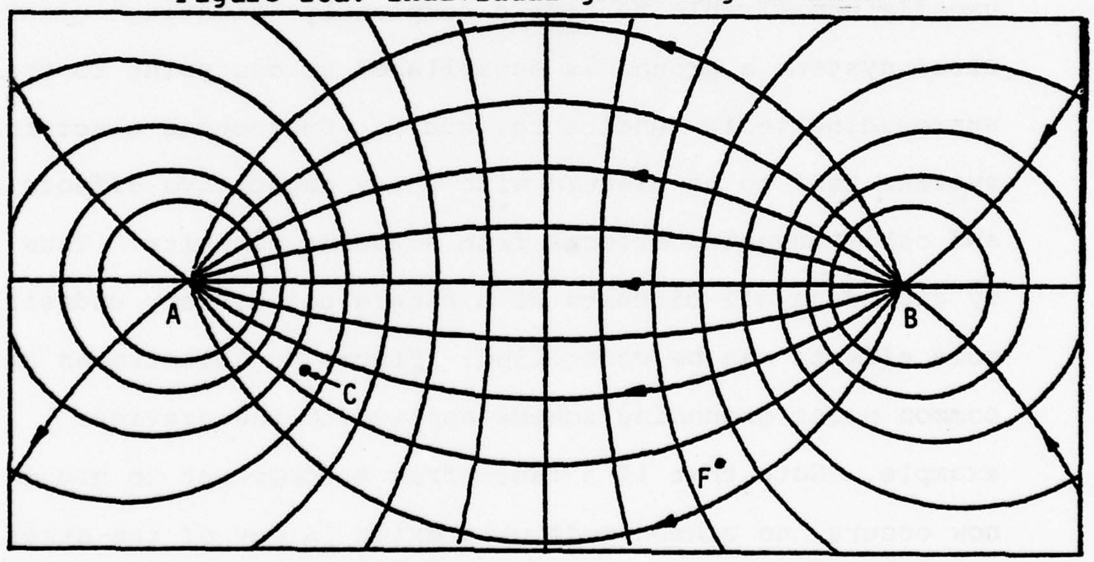


Figure 28b. Current distribution and equipotential lines for APU system (Fig. 28a)

and other design approaches are valid and useful in many cases. However, due to the severity of expected ground currents (even in the CW case), the large number of differing design philosophies involved, and the cost effectiveness of using standard components (standard logic levels, etc.), an alternate grounding scheme can be used to insure proper system operation.

A method of preventing common mode voltages due to ground currents is to use a separate ground conductor for each subsystem. Obviously, these ground conductors must be electrically insulated from each other. Because it is usually undesirable to have a completely "floating" electrical system, a ground is established at one point to the surrounding media (enclosure, etc.). Ungrounded electrical systems tend to be plagued with stray capacitive effects and other unwanted effects from adjacent circuits. Thus, by grounding all circuits at a single point, many undesirable effects can be controlled. Figure 29 illustrates the common point grounding scheme applied to the previous example. Note that if a fault from a component to ground now occurs, no common mode will exist in any of the other circuits.

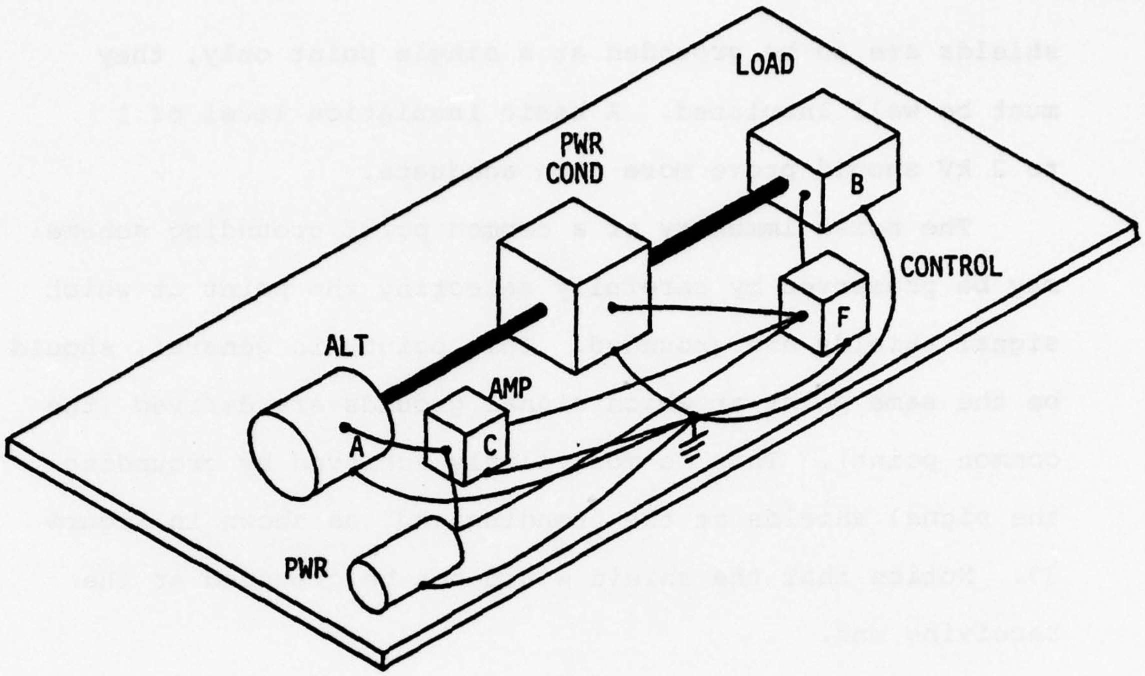


Figure 29. Common point grounding of an APU

Electrostatic shielding on high voltage equipment is usually required to reduce or eliminate capacitively induced voltages. Grounding of the shield is necessary to achieve the required shielding, and it would seem logical to ground to the same point as the return conductors. However, common point grounding of the shields is probably not as important as single point grounding of the shields. Single point shield grounding would prevent fault currents from flowing in the shielding (except of course in the case of a fault to a shield). Thus, the risk of erroneous information induced by shield current is reduced. Obviously, if the

shields are to be grounded at a single point only, they must be well insulated. A basic insulation level of 1 to 2 kV should prove more than adequate.

The noise immunity of a common point grounding scheme may be preserved by carefully selecting the point at which signal shields are grounded. This point, in general, should be the same point at which signal grounds are derived (the common point). This is most simply achieved by grounding the signal shields at the "sending end" as shown in Figure 30. Notice that the shield would not be grounded at the receiving end.

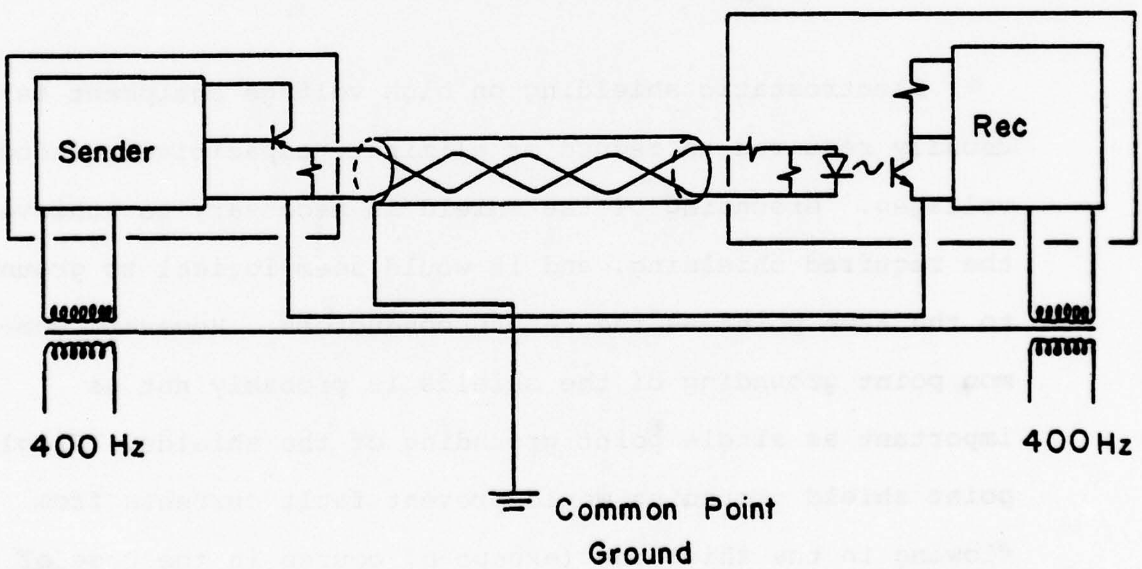


Figure 30. Shield grounding for digital inputs

By uniformly grounding the shields at the sending end, a uniform specification can be given to manufacturers of the subsystems. This will simplify technical management and control of the construction and interconnection of the many subsystems.

The termination of the signal leads at the receiving end should be in the form of optical isolators or shielded transformers to break electrical continuity and maintain the common ground integrity. The secondary or amplifier side of the terminating device would be grounded to the common point as before. Thus, all receiving lines would be isolated and the shields would not be connected at this end. It is logical that manufacturers of the receiving equipment also install the isolation devices, to reduce interface management problems. The specification to these manufacturers can thus be made uniform.

Those signals which are analog in nature can be made to meet the same specifications by incorporating a frequency modulated link (bandwidth limitations not withstanding). Thus, by using a VCO (voltage controlled oscillator) and an FM detector, analog information can be simply transmitted over the same type leads as the digital information (Figure 31). Thus, except for possible

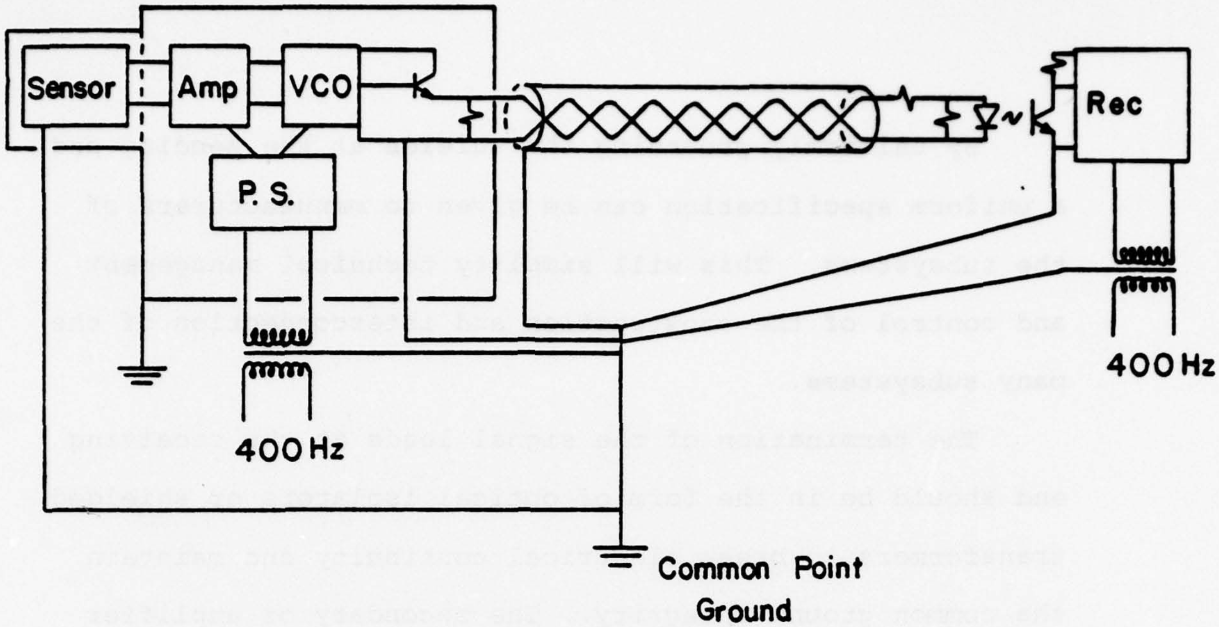


Figure 31. Shield grounding for analog system

bandwidth limitations, the grounding and shielding requirements for analog channels can be made the same as for digital channels.

For the signal leads which are physically located in the vicinity of high voltage terminals, it is advisable to observe a further precaution. Namely, the signal leads should be enclosed or contained in a conduit (Figure 32)

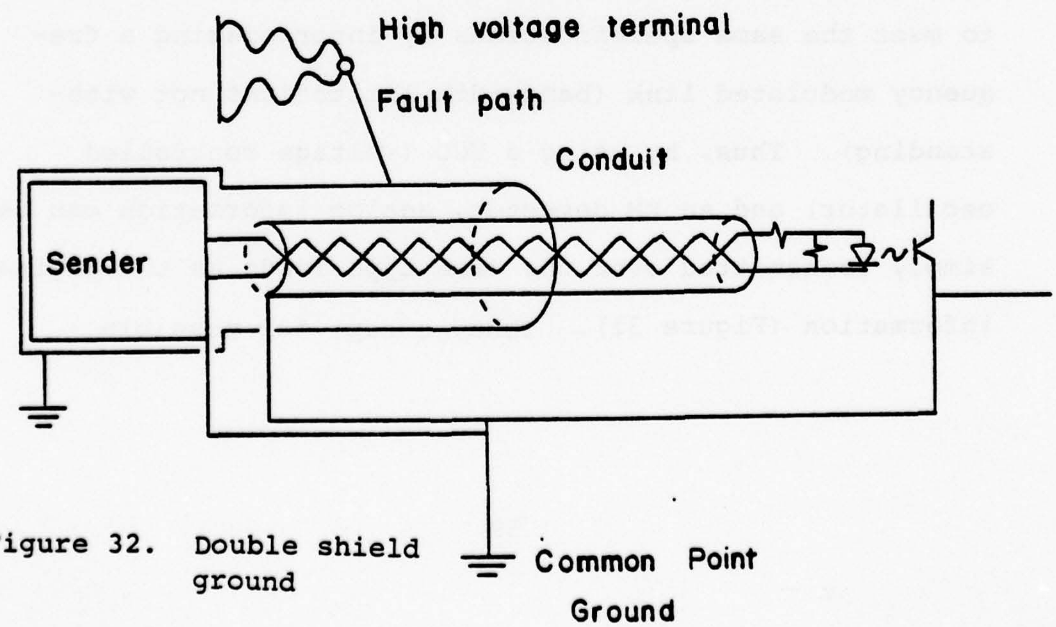


Figure 32. Double shield ground

to help insure that fault currents do not flow in signal shields. The use of conduit represents double shielding (at least for a portion of the cable runs). The conduit can be grounded to the structure, and the common point ground scheme need not be observed.

In some cases, it may be desirable to have the signal shield grounded at the receiving end. By using a doubly shielded cable, one shield can be grounded at the receiving end and the other at the sending end (Figure 33). It is not clear that this arrangement has any particular advantages and is put forward at this time only as a point for consideration.

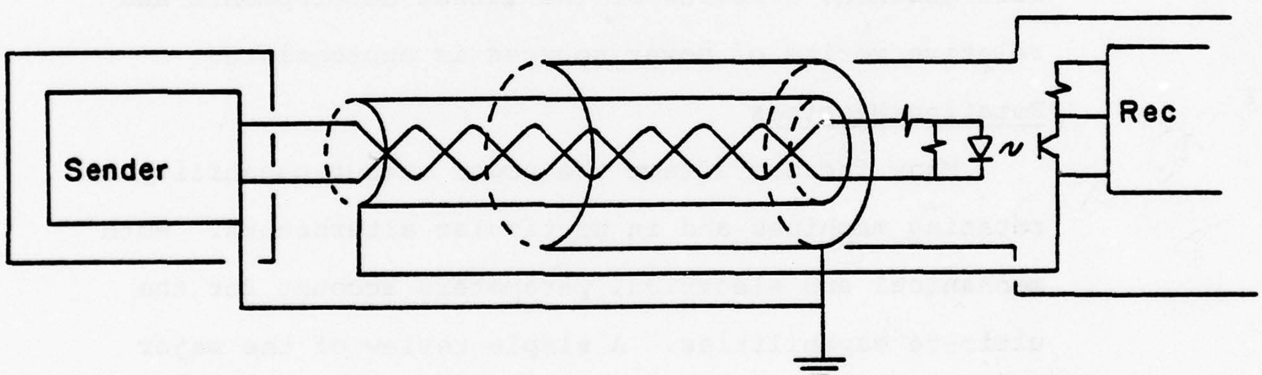


Figure 33. Double shield

SOURCE REVIEW

The requirement for lightweight and compact power sources places an extreme on material science and designer ingenuity. Each application requiring electrical power at some voltage, current, frequency, etc. will result in a point design which will not, in general, be appropriate to another application. The tendency to design a source at some power level and then adjust the output by power conditioning equipment will not necessarily result in a simpler or lighter system. Thus, a systems approach is required to achieve the most desirable result. However, because many of the future applications are not well defined, a review of the latest developments and relative merits of power sources is appropriate.

Rotating Machines

Many factors affect the power output capability of rotating machines and in particular alternators. Both mechanical and electrical parameters account for the ultimate capabilities. A simple review of the major factors will help define the reasons for development areas in alternator research.

To assist in this review, an ideal situation is postulated. Suppose that a lossless alternator is available with zero internal impedance. Thus, the mechanical input

power will equal the electrical output power. Let the electrical load consist of a resistor R. The electrical instantaneous output power, p, is then

$$p = \frac{e^2}{R} \tag{21}$$

where e is the instantaneous output voltage. A source of magnetic flux is, of course, present and the output voltage is given by Faraday's Law. By design, the flux is made to alternate in the stator windings so that if

$$\phi = \phi_m \sin\omega t \tag{22}$$

then Faraday's Law gives

$$e = N\phi_m \omega \cos\omega t \tag{23}$$

where N = the number of turns

ϕ = Magnetic flux (webers) linking the stator winding

ϕ_m = Maximum value of flux

ω = Angular rate of change due to mechanical rotation (rad/s)

t = Time(s)

Thus, the power output is given by

$$P = \frac{N^2\phi_m^2\omega^2\cos^2\omega t}{R} \tag{24}$$

and, by trigonometric substitution, the average output power P_{av} is

$$P_{av} = \frac{N^2\phi_m^2\omega^2}{2R} \tag{25}$$

Clearly, the power output is a function of the square of the magnetic excitation and the mechanical speed of the rotor. Because there is an ultimate strength of the rotor material, the rotor speed is limited by centrifugal forces. This limit can be shown to be a function of the tip speed of the rotor. Thus, even for a lossless machine, the power output is limited by the strength of the rotor material under the conditions of an arbitrary, but fixed load.

From mechanical considerations alone, the rotor tip speed can be significantly increased by tightly wrapping the rotor with a high strength composite material. This concept has been tested within the time frame of this report. Spin tests were performed and were successful in demonstrating that significantly higher speeds are obtainable over what the rotor core material could achieve alone (about a 30 percent increase). However, for a classically wound rotor, other problems may negate the apparent advance possible. For instance, the air gap will be increased because of the presence of the composite material (recall that the power output is proportional to the square of the excitation flux). Also, rotor growth due to the high speed will result in the excitation windings moving in the rotor slots causing chaffing of the insulation to say nothing of the insulation stress. A combination of permanent magnet

excitation and the filament wound rotor would seem to have great promise for significant weight and size reduction over the present state of the art of classical alternator design. A more detailed discussion on the use of permanent magnets for excitation is covered in a later section.

As previously stated, the strength of the excitation flux strongly affects power output capability. For classically designed machines, the peak flux density is limited by the saturation characteristics of the iron used in the construction. Clearly, if size and weight are to be reduced by increasing excitation strength, more exotic (and expensive) materials must be employed. Little or no effort has been made in this direction within this report period.

The results of a recent study conducted by Aeropropulsion Laboratory (APL) at Wright-Patterson Air Force Base (WPAFB) indicated that classically designed machines above the 10 MVA level are severely limited by practical considerations. The lightweight designs utilized rotors with high length-to-diameter ratios. This results in balance problems as well as the system requirement of turning motors to prevent a set or warp in the rotor shaft after operation.

Another factor which strongly affects the power capabilities of alternators is the removal of heat generated by losses. Excessive temperatures cause degradation of electrical insulation. Insulation temperature is usually the

limiting factor on the allowable temperature rise. Thus, heat management is an important design constraint and limitation. Current density in the windings is the usual figure of merit used for comparison.

For some applications where the duty cycle, etc. can be clearly defined, a point design using no cooling may result. This is especially true for short run times. The trade-off is clearly between winding mass and the complexity and weight of the required cooling equipment. Heat management techniques such as heat pipes have yet to be effectively applied to alternator design. This could result in a significant weight reduction if some means could be found to use this cooling technique. Higher current densities and thus less iron would be required for the same power output.

The output voltage required also strongly affects alternator design. Not only is electrical insulation a poor conductor of heat but it occupies space in the slots that could be used for iron or conductors. At the higher potentials, the volume occupied by the insulation becomes prohibitive so that the addition of a transformer may result in lower system weight. The APL High Power Study indicates 30 kV as an approximate upper limit for classically designed alternators with present technology. In addition, the coolant must operate in the high potential

environment further complicating the system.

An alternative to the transformer trade-off is the ring or gramme ring winding on the stator. This scheme was used in the early days of alternator development and is suggested (by Westinghouse) for use at the higher voltage levels (>30 kV). The attractive feature of the gramme winding is that the insulation is stressed to approximately phase voltage rather than line-to-line as in conventional designs. This design tends to use more conductor (see Figure 34) than conventional designs, especially for long machines.

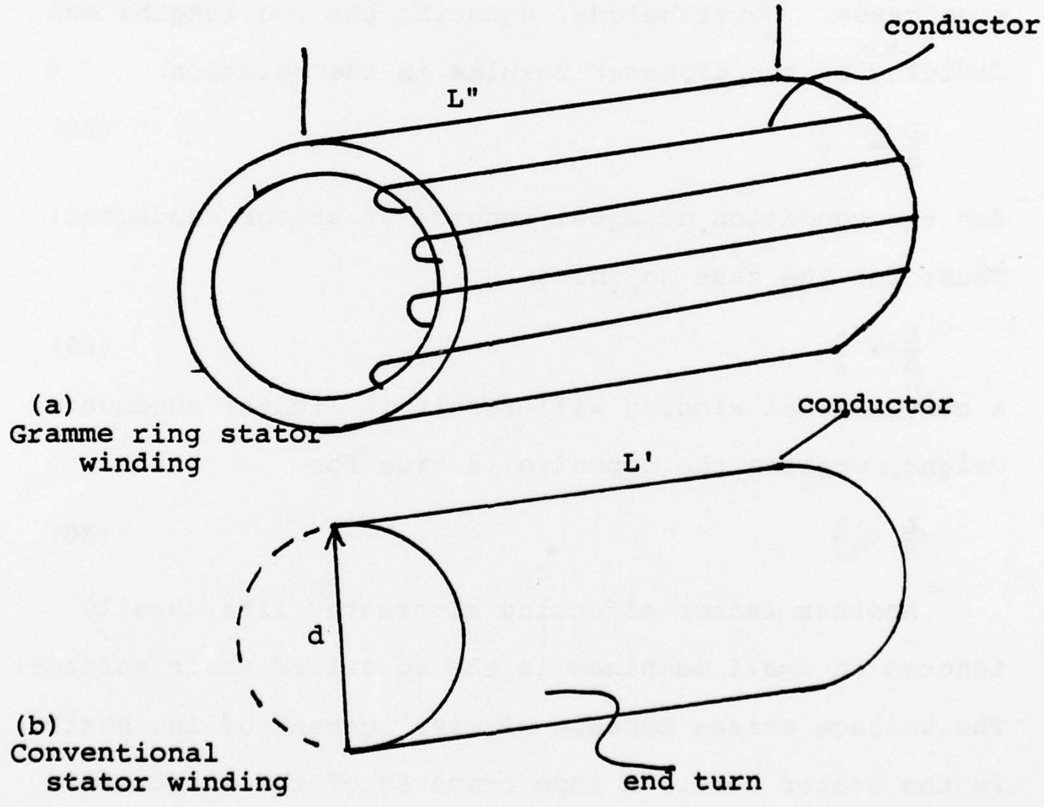


Figure 34. Alternative stator designs

A simplified calculation serves to demonstrate the cross over in conductor weight between the two concepts. Referring to Figure 34, one half the length of a stator conductor is given approximately by the formula

$$L_C = L' + \frac{\pi d}{2} \tag{26}$$

for the conventional winding and

$$L_g = 2L'' \tag{27}$$

for the gramme winding. Of course, conductor length required for bending and overlap has been ignored in both cases. This approximation is an oversimplification in some cases. Nevertheless, equating the two lengths and dividing by the diameter results in the relation

$$\frac{L}{d} = \frac{\pi}{2} \tag{28}$$

for the condition of equal lengths of stator conductor. Thus, for the case in which

$$\frac{L}{d} > \frac{\pi}{2} \tag{29}$$

a conventional winding will result in lighter conductor weight, whereas the opposite is true for

$$\frac{L}{d} < \frac{\pi}{2} \tag{30}$$

Another factor affecting alternator life usually ignored in small machines is the so called shaft voltage. The voltage arises because of misalignment of the shaft in the stator bore. A loop composed of the shaft, end

bells, back iron, and the shaft will not have equal voltages induced so that a current will flow in the shaft and, typically, through the bearings. This current will cause pitting of the bearing surfaces and reduce bearing life. As power output levels increase, pitting will become an increasingly important consideration. If recognized, pitting of bearing surfaces can be easily controlled.

Permanent Magnet Alternators

Permanent magnet alternators offer a possible solution for the stringent requirements for airborne APUs' namely, the promise of light weight, fast start, and little or no cooling. Because the excitation is always applied (permanent magnets), poor voltage regulation and control as well as fault control are limitations which may initially detract from the desirability of this type alternator, however, a careful analysis of the failure modes and appropriate design of the output control circuits may alleviate most of the objectional features.

Classical machines operate with air-gap field strengths in the range of 1 to 1-1/2 Tesla. Permanent magnet materials which can achieve similar flux densities have been available for many years. However, because of the demagnetizing effect of load current and especially fault currents, permanent magnets were not practical for use in large machines. Poor voltage regulation and the possibility of the loss of excitation after a fault have limited permanent magnet machines

to small, special purpose applications.

Recent developments in magnetic materials have led to the development of magnets with very high energy products. New rare earth-cobalt materials are commercially available with energy products of 22 megagauss-oersteds and show promise of an ultimate of 50 megagauss-oersteds.

The application of this new material to lightweight, high-power alternator design should result in a new class of power sources.

Efforts to apply the new magnets for alternator excitation are currently underway at the Garrett Corp., California. The target specification for this study is 0.045 kg/kW (0.1 lb/kW) and the initial base line design is for a 1 MVA, thermal lag machine. The tip speed was initially set at 625 ft/s and the stator current density at 20 kA/in^2. Higher tip speeds and current densities are indicated for the high power designs to be included in this study. A beryllium copper ring will serve as a damper winding and to structurally contain the rotor.

Several important aspects of design limitations are scheduled in the critical component testing portion of the study. Principally, the temperature and stress on magnet behavior are to be evaluated. Because permanent magnet materials are known to be structurally weak and the residual magnetism is sensitive to temperature, these two factors may be the limiting ones in the design. Although there is no

heat dissipated in the rotor due to excitation, some heating nevertheless occurs because of windage and losses occurring in the damper due to variations in the air gap flux. Load variations, commutation of rectifiers, and tooth reluctance are factors leading to rotor heating. The initial estimate on the magnitude of these effects is approximately 5 percent reduction in magnetism due to the stress. If the two effects are cumulative, a total change of 10 percent can be expected.

Because the rotor will require some cooling, a simple cooling scheme is employed wherein heat is conducted to the shaft by aluminum fins inside the rotor and removed by a coolant circulated through the shaft.

The voltage regulation is expected to be approximately 20 percent. As there is no direct means of adjusting alternator output, an active external circuit must be employed to achieve the required regulation. Although this may be viewed as an objection, a closer examination will reveal that external regulation is required by most of the applications anyway.

Regardless of the power source, rapid and responsive voltage control is required. Also, because of the likelihood of load faults, rapid removal of power is a stringent requirement. Neither of these requirements can be met with field control alone. Thus, even for classical alternators,

external circuits are required and field control employed only for long term regulation and ultimate fault control. By proper design of the external circuits Silicon Controlled Rectifiers (SCR) bridges, etc. the response of the permanent magnet machine can be made to be equivalent to that of a classical machine. Also, because of the limited amount of excitation available with permanent magnets, extreme over-voltages will not occur. For a fault internal to the machine or in the external circuits, no means is available to limit fault energy except by stopping the rotor. Since the contemplated designs are fast start, possibly a fast stop feature could be incorporated to limit fault energy.

The permanent magnet alternator is an operationally attractive concept. Cooling requirements are greatly reduced and, with magnetic materials improvement, very light weight is possible. Because preliminary designs indicate a high frequency output, a reduction in power conditioning component size is possible. The only significant operational objection is ultimate fault control. Utilizing fast stop, control fuses, etc. can help alleviate this objection. For internal, electrical faults, the mission is over and a restart is out of the question.

Superconducting Alternators

Significant advances have recently been made on the superconducting alternator. A superconducting rotor has been spin tested and excited at 110 percent rated speed.

A complete 10 MVA machine is in the final stages of assembly and limited testing is planned before the end of the year.

The principal weight reduction with a superconducting rotor results from the significantly higher current densities possible. This results in very high field strengths (>4 Tesla) and no iron is required (except for structure) in the rotor. The field at the stator windings is approximately the same as in classical machines. Clearly, a large air gap is present resulting in a low internal impedance. Reaction to short circuits will result in very high fault currents, and elaborate shielding of the superconducting (s.c.) field is required to prevent the field from going "normal."

The tip speeds possible with a superconducting rotor are, and will probably remain, significantly lower than other designs because of the extreme sensitivity of present superconducting materials. Small movements and changes in the mechanical stresses may cause the conductors to go normal. Current designs limit the tip speed to 420 ft/s as opposed to the 600 to 700 ft/s common in classical designs.

Also, because there is no iron to guide or control the flux distribution, the space between poles must be kept large compared to the bore air gap to prevent excessive leakage flux. Thus, the number of poles is limited if acceptable stator flux is to be maintained. The combination

of tip speed and pole limitation combines to limit the output frequency possible with this concept.

Proper evaluation of a superconducting alternator for airborne use is more difficult than the classical or permanent magnet concepts. Considerable ground support equipment is required because of the cryogenic cooling requirements, and flight time is limited by the on board helium supply. Recent studies indicate that the volume of the on board helium supply for a 12 hr ready time is approximately equal to the alternator volume itself.

In many respects the superconducting alternator is similar to the permanent magnet machine. Voltage control can be achieved, but rapid, responsive control is unlikely because of the nature of the superconducting field and because of size limitations of the charging supply. Present designs of the charging supply require aircraft power and become large for rapid excitation and control. Ultimate fault control requires quenching the field, making a restart difficult due to heating of the superconductor.

Further, for some applications, the limited frequency capabilities place more restraints on the power conditioning equipment. Clearly, for direct current applications the filter required is larger than for higher frequency supplies. This places additional constraints on energy diverters in the load protection circuits. These disadvantages may be

more than offset in some applications by the fact that the superconducting generator is well suited to generate high voltages; i.e., it has a low length to diameter ratio (L/d). (See previous discussion on gramme ring winding). Also, the use of a larger number of phases is not precluded.

The results of the high power study conducted by APL indicate that the specific weight for the alternator ranges from 0.0985 to 0.055 lb/KVA, and for the complete system, 0.0998 to 0.132 lb/KVA depending on the power level and output voltage. Specific volumes range from 0.8 to 2.2 ft^3/MVA. All designs utilized open cycle cooling for the rotor, and no heat sink designs were included.

Magnetohydrodynamic Generators

The MHD generator has been studied in great detail, but no great advances have been made that indicate a truly important advance in the utility of this concept. Present research in various laboratories indicate that significantly higher power densities are possible than was attempted in domestic work, but the problems associated with operation of an MHD machine in a confined space have only recently been addressed. Shielding of the required magnetic fields is an important problem for airborne application. Current work to shield the 4 to 5 Tesla fields involves active or image shielding; i.e., conductors are placed appropriately around the main magnet to cancel external fields in an active sense.

This concept allows attractive weight projections to be made. Exotic liquid fuels and arrangements of solid fuel rocket motors have been investigated to perform a number of start-run cycles, and these systems are attractive from a projected weight/volume point of view.

Recent efforts involve a generator utilizing JP-4, LOX as the fuel. The generator is configured as a Hall generator. Several problems have delayed the testing of this machine, and the results will not be available within this report period. The projected output is approximately 1 MJ/kg fuel with a thermal efficiency of approximately 10 percent. Present plans call for testing of the combustor, flow channel, etc., and no electrical power will be extracted because a magnet will not be available.

Since the MHD generator is somewhat limited in design flexibility due to sensitivity of the design voltage to power output requirements, it is probable that power conditioning equipment will be a requirement. Inverters are likely to be required and may operate at 30 to 50 kHz. Switches required are probably within the state of the art but, as in most switching schemes, the capacitors again are a problem.

In order to achieve favorable weight and volume, the MHD machine is loaded to achieve maximum power transfer. Thus, the electrical efficiency is 50 percent and the voltage regulation is also 50 percent. Because the machine is

designed for this condition, an open circuit will very likely result in an arc in the channel. This condition is to be avoided by placing a short on the machine terminals. Normal conditions would be reestablished by seed control. Operationally, the shorting network is yet to be defined.

The flight weight machine requires a superconducting magnet. Four to six Tesla fields are required over a considerable volume. The consequences of interaction of the mechanical portion of the MHD generator in this high field are yet to be determined. The reaction of the magnet to a channel arc is not clear at this time. All in all, the MHD generator is an attractive source, but insufficient evaluation has been made of the system effects to make a realistic evaluation of the desirability of this source.

Batteries

Batteries hold great promise for truly light weight and high density power sources. A simple evaluation for the weight of the reactants required to produce a given number of watt-seconds gives ridiculously small numbers. The difficulty comes from the weight and volume of the electrolyte and the separators. Recent research on new separators and bipolar electrodes for silver-zinc was only partially successful. The bipolar electrodes appear

to be practical, but the separators failed to prevent silver migration. The migration of the silver greatly reduces the shelf life and the rechargability of the battery. Approximately 70 percent of the weight of the silver zinc battery is contained in the electrolyte. Obviously, development of a new separator that reduces the amount of electrolyte required would have significant impact on battery sources. Apparently, no new research is planned for this power source.

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